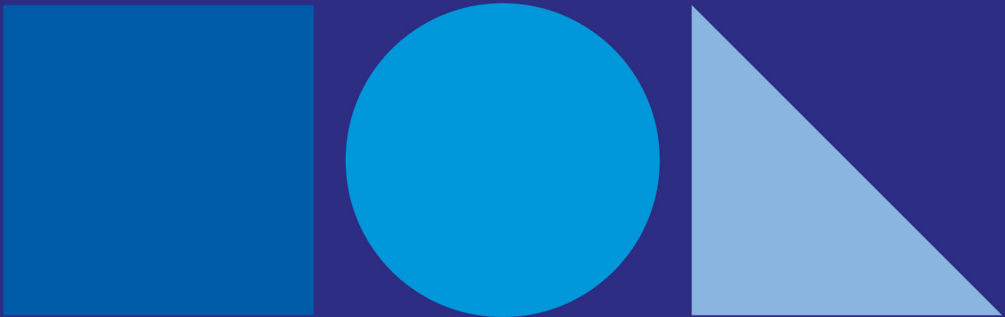


Resonant Tunneling Diode Photonics

Devices and applications

Charlie Ironside
Bruno Romeira
José Figueiredo



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*This book is dedicated to all our past and existing co-workers who have helped
advance resonant tunneling diode photonics.*

*Bruno dedicates the book to Raquel, Carolina, Milene, and Adelaide.
José also dedicates the book to Rita, Inês, and Cristina.*

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Preface

Resonant tunnelling diodes (RTDs) exhibit two main features at room temperature when compared to other semiconductor devices, wideband negative differential conductance (NDC) and an extremely high operating frequency reaching 1.92 THz. The wideband NDC leads to electrical gain and when DC polarized in the NDC region the RTD produces a radio frequency (RF) signal whose frequency is determined by its intrinsic capacitance and by the equivalent inductance of the connecting circuitry. The high operating frequency, on the other hand arises from the very thin double barrier quantum well (DBQW) structure, which makes the RTD the fastest pure solid-state electronic device operating at room temperature. These unique features make it possible for the RTD to operate as a voltage controlled electronic oscillator (VCEO), providing an interesting solution for high-speed communications. Moreover, the portion of its current–voltage (I – V) characteristic exhibiting negative differential conductance (NDC) provides the electrical gain needed to implement very simple high frequency electrical oscillators and countless other electronic and optoelectronic devices with new functionalities.

A wide number of optoelectronic devices and circuits based on III-V semiconductor compound materials can benefit from the integration with RTDs, taking advantage of the NDC produced by DBQW structures. Such optoelectronic devices include laser diodes, semiconductor optical amplifiers (SOA), photo-detectors and electro-absorption modulators. Interestingly, in the past years resonant tunnelling diode electronics and optoelectronics high-speed technology has already demonstrated wide-spread potential for disruptive applications which include THz emitters and detectors for imaging and communications beyond 5G, single-photon switches and detectors and mid-IR detectors for sensing, and more recently neuromorphic functionalities for neuromorphic computing and artificial intelligence systems.

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Author biographies

Charlie Ironside



Professor Charlie Ironside is a professor in the Department of Physics and Astronomy at Curtin University (Western Australia). He received his PhD from Heriot-Watt University (Edinburgh) and in 1998 was appointed Professor of Quantum Electronics at the University of Glasgow. He has over 30 years of experience in semiconductor optoelectronics research and in particular, micro-fabrication of photonic components. He is a Fellow of the IET, a Fellow of the Institute of Physics, and a senior member of IEEE. Along with hundreds of papers, he has co-authored five patents. In 2012 he won the knowledge transfer partnership Scotland award.

Bruno Romeira



Bruno Romeira received a PhD degree (summa cum laude) in physics and the European PhD degree from the University of Algarve, Faro, Portugal, jointly with the University of Glasgow, UK, and the University of Seville, Spain, in 2012. From 2013 to 2014, he was a Postdoctoral Fellow with the Microwave Photonics Research Laboratory, University of Ottawa, Canada. From 2015 to 2017, he was a Marie Skłodowska-Curie Research Fellow with the Applied Physics Department, Institute for Photonic Integration, Eindhoven University of Technology, The Netherlands. He joined the International Iberian Nanotechnology Laboratory (INL), Braga, Portugal, as a Marie Skłodowska-Curie COFUND Research Fellow in 2017, where, since 2019, he has been a Staff Researcher and a Coordinator of the EU-H2020-FET-OPEN Project ChipAI. His research cuts across several disciplines in applied physics and engineering, which include semiconductor physics, quantum nanoelectronics, low-dimensional nanostructures, and nanophotonic and neuromorphic devices.

Dr Romeira was a recipient of the Young Researchers Incentive Programme Award from the Calouste Gulbenkian Foundation, Portugal, in 2009, the 2011 IEEE Photonics Society Graduate Student Fellowship from the IEEE Photonics Society, USA, and the Best PhD Thesis in optics and photonics in Portugal in 2012 Award from the Portuguese Society of Optics and Photonics. Dr Romeira is a member of the Optical Society of America and of the IEEE Photonics Society.

José Figueiredo



José Figueiredo received a BSc in physics (optics and electronics) and a MSc in optoelectronics and lasers from the Universidade do Porto, Portugal, in 1991 and 1995. He did his PhD work at the University of Glasgow, Scotland, on the optoelectronic properties of resonant tunnelling diodes, enrolled in a joint programme between the Universidade do Porto and the University of Glasgow.

Then he moved to the Universidade do Algarve, Portugal. He joined the Physics Department of the Faculty of Sciences of the University of Lisbon in 2017. He is member of the CENTRA research unit (tecnico.ulisboa), Portuguese Society of Optics and Photonics, European Microwave Association, Optical Society of America, and IEEE Photonics Society.

In 2001, he co-won the Scottish Proof of Concept Fund for transferring knowledge from university research to industry for commercial exploitation. He co-founded the start-up Essient Photonics, a University of Glasgow spin out company specialized in optical modulators and photodetectors that take advantage of the properties of resonant tunnelling diodes. He has published over 70 journal and conference papers, and co-authored four patent applications. His research interests include nanoelectronic and optoelectronic devices for terahertz, wireless, radio-over-fibre, neuromorphic, reflectometry, and sensing applications.

Resonant Tunneling Diode Photonics

Devices and applications

Charlie Ironside, Bruno Romeira and José Figueiredo

Chapter 1

Introduction

1.1 Introduction

This book brings together two broad themes that have generated a great deal of interest and excitement in the scientific and technical community in the last 100 years or so. These themes are quantum tunnelling and nonlinear dynamical systems and, in the following chapters, we apply them to nanostructured solid state heterostructures operating at room temperature to gain insight into novel photonic devices, systems and applications.

Throughout the history of microelectronic solid state devices, charge carriers have mostly been treated as classical particles, essentially obeying Newton's laws of classical mechanics—this is the fundamental assumption behind Boltzmann's transport equation and the drift-diffusion model of electronic transport in solid state devices [1, 2]. The wave nature of the electron and how it interacts with the period lattice of the crystal is incorporated into this theory by the effective mass approximation. The electron is treated as a particle that obeys classical mechanics but which has a mass that is different from the free space mass. This semi-classical approximation works because the wave nature of the electron is masked by scattering; that is the electron bumps into entities such as lattice ions, impurity ions and other electrons and these collisions randomly reset the phase of the electron wavefunction. So the electron wavefunction is incoherent on a scale larger than the average distance between collisions—called the phase breaking distance [3]. This phase breaking distance is larger than the lattice spacing which is typically around 0.5 nm—that is why the effective mass approximation works. It turns out that, at normal working temperatures, around room temperature, and in typical semiconductors, the phase breaking distance is less than around 10 nm or so [4]. Another way to express this is to say that phase breaking is a decoherence effect and if we are to observe, and indeed make use of, quantum mechanical tunnelling effects then we have to make structures that are small enough that the coherence of the electron wavefunction is preserved and so typically the critical dimensions of quantum

tunnelling devices, in room temperature semiconductors, are approximately 10 nm. This is the scale of the tunnelling devices we discuss in this book—the so-called mesoscopic scale [4]. To quote Alain Aspect ‘between the scale of a single atom and the macroscopic world, one finds the mesoscopic scale, where it is the object itself, and not only the material of which it is made, that needs to be described by quantum mechanics’ [5].

Using the tools of semiconductor growth and nanofabrication we can make mesoscopic devices that utilise the electron wave nature to make new and interesting charge carrier transport and optical properties and produce macroscopic, room temperature, nonlinear current–voltage (I – V) characteristics that exhibit negative differential conductance; that is, in some part of the I – V curve as the voltage increases the current decreases. This is achieved by utilising quantum mechanical tunnelling through a resonant state. The devices are two terminal devices and are thus referred to as diodes, and so we have the resonant tunnelling diode (RTD). The RTD [6] is a mesoscopic device—these devices are deliberately designed to employ the wave nature of the electron to achieve device benefits.

The importance of scale and the transition between the appropriateness of a quantum description of charge carrier transport compared to a classical description is neatly illustrated in [7].

The wave nature of electrons in solids has become of increasing importance because the drive to make more compact and faster devices has led to diminishing dimensions of important features in solid state electronics. The size of a MOSFET (metal oxide field effect transistor) gate has shrunk down to less than around 10 nm (at the time of writing in 2019) and is continuing to shrink. It can be anticipated that the wave nature of electrons and quantum tunnelling will play an important part in the MOSFET operation because the dimensions are approaching the typical phase breaking distance in room temperature silicon. Already tunnelling theory (see, for example, [8]) is used to describe leakage currents in MOSFET gates which is caused by quantum tunnelling, the so-called direct tunnelling and Fowler–Nordheim tunnelling [9].

The feasibility of fabricating mesoscopic devices was greatly facilitated by epitaxial growth of semiconductor heterostructures. Techniques such as molecular beam epitaxy (MBE) and metal organic chemical vapour deposition (MOCVD) can accurately deposit on a substrate crystalline epilayers of semiconductor alloys with thickness as small as one monolayer (~0.28 nm). The GaAs/AlGaAs semiconductor alloy was the system that was first to be investigated and the application was in photonics leading to the remarkable development of the double heterostructure semiconductor laser [10]. These growth techniques allowed the development of mesoscopic photonic devices that deliberately employed the electron wave nature of the electron to achieve improved functionality; these include the quantum well laser and the quantum cascade laser.

1.2 Quantum tunnelling devices

In terms of semiconductor devices, the so-called Esaki diode [11] was the first device that employed the quantum tunnelling effect to achieve a nonlinear current–voltage curve. The Esaki diode is similar to a p – n junction diode but with highly doped p and n regions that give rise to a small depletion width on a scale that allows tunnelling, that is, the depletion width is smaller than the phase breaking distance. The Esaki diode was the first clearly demonstrated mesoscopic electronic device operating at room temperature and its discovery resulted in Esaki being awarded the Nobel prize in 1973. The same year as the prize was awarded to Josephson for the discovery of the superconducting metal–insulator–metal device, the Josephson junction, that relies on quantum tunnelling for its operation [12].

In this book we are concerned with resonant tunnelling diode (RTD) devices which belong to a class of semiconductor heterostructure devices termed mesoscopic devices. In these devices the dimensions in the direction of current flow are smaller than the so-called phase breaking length [3] of the charge carrier in the device. Phase breaking is caused by the various scattering processes that a charge carrier is subject to in a semiconductor, as for example phonon collisions or electron–electron scattering—it can be regarded as the distance over which the electron wavefunction remains coherent. Further, if quantum effects are to play an important role in these devices, the temperature is required to be below the Thouless temperature [3, 13]—*the temperature at which thermal averaging introduces an uncertainty of π radians in the electron's phase*. The phase breaking length determines the critical dimensions in the RTD device.

Since in this book we are covering electronic mesoscopic structures combined with photonic devices, it is worthwhile pointing out some of the striking analogies between optics and quantum devices; the unifying theme is wave mechanics. In optics the design of thin film devices such as antireflection coatings proceeds via transfer matrix methods [14] and in a similar way these methods can be applied to electronic quantum devices such as the RTD [15–18]. In electronic wave mechanics, the electron wavefunction decays as it penetrates through a barrier and the optics analogy is the evanescent wave [19] that is present in the low refractive index part of optical waveguides and generally in the low refractive index part of optical devices that employ total internal reflection.

1.3 Negative differential conductance

There are many physical mechanisms that give rise to negative differential conductance (NDC) and thereby many devices that display NDC characteristics ranging from Gunn diodes [20, 21] to gas discharge tubes [22]. In this book we are concerned with NDC arising from quantum tunnelling. How quantum tunnelling leads to NDC is discussed in chapter 2. What we mean by NDC is best explained with reference to figure 1.1 that shows a typical nonlinear current–voltage characteristic for an RTD.

When DC is biased to the appropriate voltage an NDC device exhibits a region of negative conductance (green shaded region in figure 1.1), that is, as the voltage is

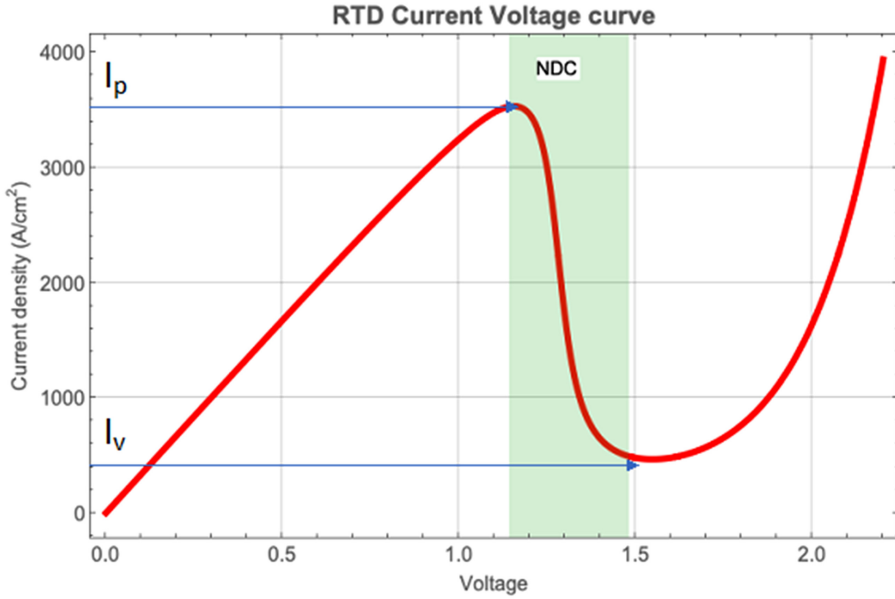


Figure 1.1. The fit to the current–voltage (I – V) curve of a InGaAs/AlAs RTD (taken from [23])—the NDC is exhibited for voltages in the range ~1.1 to 1.5 V—marked on the curve. The peak (I_p) and valley currents (I_v) are shown.

increased the current decreases. In the NDC region of the I – V curve the device has electrical gain, i.e. it generates power rather than absorbs power. When biased to the appropriate voltage, in this example around 1.1 V, in figure 1.1, the NDC region has electrical gain. The amount of electrical gain is largely determined by the ratio of the peak (I_p) to valley currents (I_v).

Note: Since conductance and resistivity have a reciprocal relationship the resistance of the devices will be negative in the NDC region and often the NDC region is referred to as the negative differential resistance (NDR) region.

Not apparent from figure 1.1, is the speed at which the device can switch between the peak to valley currents. It turns out that the RTD can have a very large gain bandwidth due to the speed of quantum mechanical tunnelling switching which is very fast (<1 ps). In fact, electronic oscillators based on RTDs have operated at up to 1.93 THz [24]. Therefore, a great deal of the research interest is centred on RTDs because of their high-speed operation.

1.4 Nonlinear dynamics

Early in the history of the Esaki tunnel diode it was realised that the nonlinear I – V exhibiting NDC could be used as an analogue of other nonlinear dynamical systems such as simulating a nerve axon [25], see figure 1.2 which forms part of the

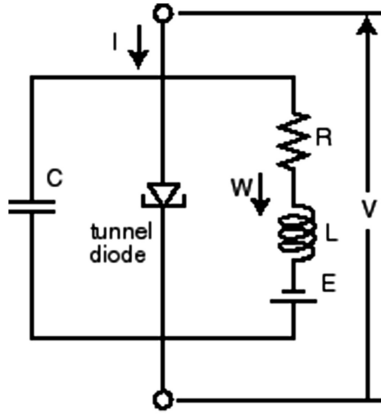


Figure 1.2. The circuit diagram of the tunnel-diode nerve model [25].

FitzHugh–Nagumo model [26] of nerve operation. This model is a simplified version of the well-known Hodgkin–Huxley model describing the behaviour of ion channels that permeate the cell membrane of the squid giant axon. Hodgkin and Huxley were awarded the 1963 Nobel Prize in Physiology or Medicine for this work.

Nagumo *et al* [25], inspired by how the nonlinear operation of nerve cells could be explain by the Bonhoeffer–van der Pol (BVP) model [27], demonstrated how the circuit illustrated in figure 1.2 could model the nerve cell activity. Further, they discussed the conditions under which the circuit operated as a Liénard-type oscillator and much later the Liénard-type oscillator approach was also applied to the RTD–laser diode (RTD–LD) [28].

To keep in accordance with the van der Pol oscillator theory, Nagumo *et al* [25] used the equation of a third order polynomial equation, $F_N(V)$, to fit the I – V characteristics given by:

$$F_N(V) = i_0 - \frac{1}{\rho} \left((V - V_0) - \frac{(V - V_0)^3}{3K^2} \right) \quad (1.1)$$

where ρ and K are adjustable parameters with the dimensions of resistance and voltage, respectively, and $i_0 = J_0 \times \text{Area}$ and V_0 are shown in figure 1.3.

With this equation (1.1), Nagumo *et al* were able to show the correspondence between the van der Pol oscillator and the Esaki tunnel diode. Although, as we can also see from figure 1.3, the third order polynomial is not an exact fit to the type of current–voltage curve typically observed with RTDs, it can still be used to describe the key nonlinear dynamic properties of these oscillators (figure 1.4).

Since the original work in the 1960s, an extensive panoply of theoretical tools associated with nonlinear dynamics has been developed—for a general introduction see, for example, [29]. These tools have yielded considerable insight into the operation of both electronic and optoelectronic systems that are dominated by the nonlinear I – V of RTD components, where phenomena such as stochastic resonance [30], chaos and synchronisation have been observed [31]. Interestingly, the research

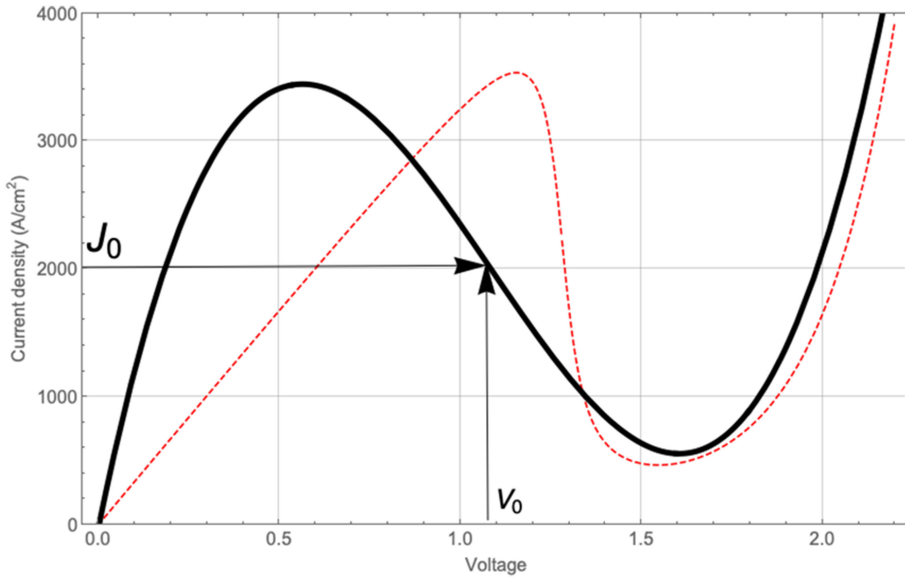


Figure 1.3. Black line: current–voltage curve of a tunnel diode fit with third order polynomial equation (1.1). Red dashed line: a comparison with the RTD device (adjustable parameters used to make comparison $V_0 = 1 \text{ V}$, $J_0 = 2000 \frac{\text{A}}{\text{cm}^2}$, $\rho = 2.4 \times 10^{-4} \Omega$ and $K = 0.52 \text{ V}$).

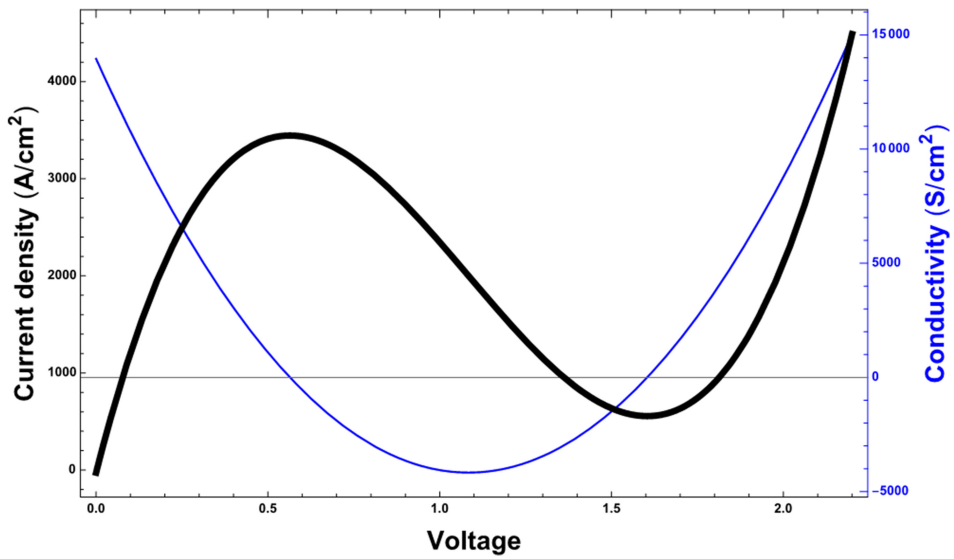


Figure 1.4. Black line: the Nagumo (van der Pol) model of tunnel diode current–voltage curve. Blue line: the differential with respect to the voltage of the current–voltage curve to give the conductance—it shows the negative differential conduction (NDC—below zero) region.

has turned full circle and now the RTD is being used to build systems that simulate neural computing [32–34].

In this book we will trace out the history of the association between nonlinear dynamics and resonant tunnelling devices and how it can be used to gain insight into the static and dynamic properties of RTD-based photonic devices and systems.

1.5 Optoelectronic integrated circuits (OEICs)

Optoelectronic technology has been inspired to emulate silicon microfabrication integrated circuits to develop technologies that bring the benefits of micro and nano integration to optoelectronic devices and systems. These benefits include high-speed compact devices with greatly enhanced functionality. Two main themes have emerged that are loosely defined as *integrated optics* where entirely optical functionality is integrated onto a single chip (see, for example, [35, 36]) and *optoelectronic integrated circuits* (OEICs) where electronic functions such electronic amplification or logic are combined with optical functions on the same chip (see, for example, [37, 38]).

This book covers how quantum mechanical tunnelling in III–V semiconductors can be employed to achieve OEIC functionality [39] in several photonic devices such as modulators (chapter 3), detectors (chapter 4) and lasers (chapter 5) and how these new tunnelling OEICs can be combined in innovative systems such as neuromorphic (brain-inspired) photonic computing (chapter 6) and optoelectronic oscillators (chapter 7).

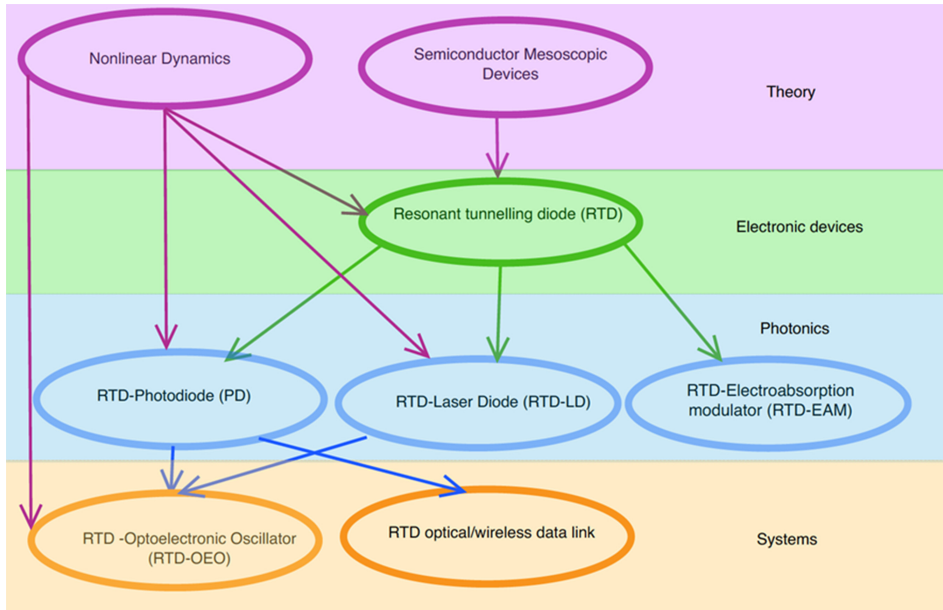


Figure 1.5. The RTD photonics map showing the relationships between theory, electron devices, photonics and RTD photonic based systems.

1.6 Outline

The major themes covered in this book and the connections between them are illustrated in figure 1.5. We cover the electronic and optical properties of RTDs (chapter 2), and their nonlinear dynamic characteristics (chapter 6) with an emphasis on how they are employed to achieve novel RTD photonic devices. This includes the resonant tunnelling diode—electro-absorption modulator (RTD—EAM) presented in chapter 3, the resonant tunnelling diode—photodetector (RTD—PD) presented in chapter 4, the resonant tunnelling diode—laser diode (RTD—LD), shown in chapter 5 and the resonant tunnelling diode—optoelectronic oscillator (RTD—OEO) presented in chapter 7.

Based on the research of many groups world-wide and focusing particularly in our work in the last 20 years, the aim of this book is to provide a background for the wide range of new opportunities emerging for RTD-based photonic devices and systems.

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Resonant Tunneling Diode Photonics

Devices and applications

Charlie Ironside, Bruno Romeira and José Figueiredo

Chapter 2

Resonant tunnelling diode—electrical and optical properties

2.1 Introduction

In this chapter, we cover the electrical and optical properties of resonant tunnelling diode (RTD) based structures which are relevant to understanding the photonic applications of these devices. As described in the previous chapter RTDs are semiconductor devices and much of their optical and electrical properties are derived from their semiconductor nature. We start this chapter assuming that the reader has a basic understanding of these properties such as can be obtained from textbooks on semiconductor optoelectronics, see, for example [1].

2.2 Differential negative conductance

Negative differential conductance (NDC) can be observed by measuring the non-linear current–voltage (I – V) characteristic for an RTD as is shown in figure 2.1. This corresponds to the key electrical characteristic employed in RTD photonic devices.

When DC is biased to an appropriate voltage an NDC device exhibits negative conductance, that is, as the voltage is increased the current decreases. In the NDC region of the I – V curve the device has electrical gain, i.e., it generates ac power rather than absorbing power. In the example of figure 2.1, when biased around 1.1 V the device has electrical gain. In the case of electrical feedback, via for example parasitic components (e.g. inductor line) and in the situation where the gain of the system overcomes the losses, the RTD will oscillate. The RTD can also be embedded in electric circuits, where oscillation is completely suppressed at all frequencies, but the circuit still provides linear or nonlinear electrical gain.

One of the really exciting aspect of RTDs is that electronic oscillators based on RTDs have operated at up to 1.93 THz. *This frequency is the highest frequency reported for a room-temperature electronic single oscillator to date* [3]. Therefore, RTDs exhibit the largest gain bandwidth of any room-temperature electronic device.

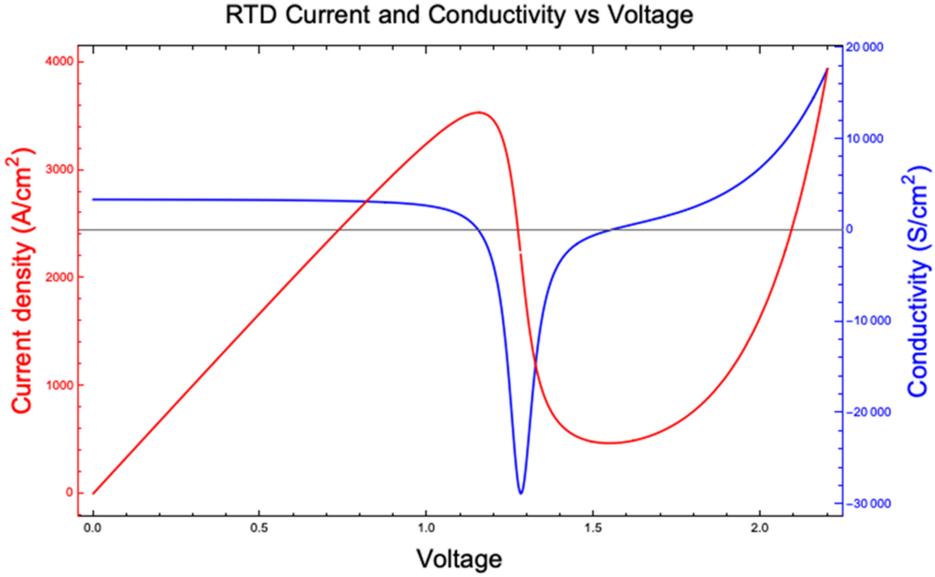


Figure 2.1. Red line: the fit to the current–voltage (I – V) curve of a InGaAs/AlAs RTD (taken from [2]). Blue line: the differential (with respect to V) of the I – V curve to give the conductivity—the NDC region is evident for bias voltages >1.2 V.

The oscillation frequency is determined by a mixture of the physical mechanisms underlying the NDC and the nature of the electrical feedback. In the RTD the speed of quantum tunnelling mechanism underlying the NDC is fundamentally limited by the dwell time or resonance time associate with the resonance state. The dwell time is the average time the electron is confined by the barriers in the quantum well and is typically less than 1 ps. From the oscillation result, this time can be less than ~ 0.5 ps and the device, therefore, has a large intrinsic gain bandwidth. Consequently, RTD oscillators can exhibit the highest oscillation frequencies. The relative contributions of parasitic and fundamental contributions to the limit of the gain bandwidth can be difficult to disentangle but it is currently the view [3] that the oscillation of RTD oscillators can be extended above 2 THz securing their position as the fastest electronic device operating at room-temperature.

The origins of the RTD I – V curve and the NDC lie entirely in the quantum effect of tunnelling through two energy barriers separated by a layer with a thickness less than the electron phase breaking distance. Therefore, to enable the quantum tunnelling effect the dimensions of the epitaxial layers that make up the RTD have to be small enough to keep the electron wavefunction coherent as is described in chapter 1 in the case of mesoscopic devices. An example of a layer structure of an RTD in the InGaAs/AlAs material system is illustrated in figure 2.2.

The explanation of how the RTD structure gives rise to the nonlinear I – V curve in figure 2.1, starts with figure 2.3 which shows the conduction band profile of a generic double barrier RTD with zero applied voltage.

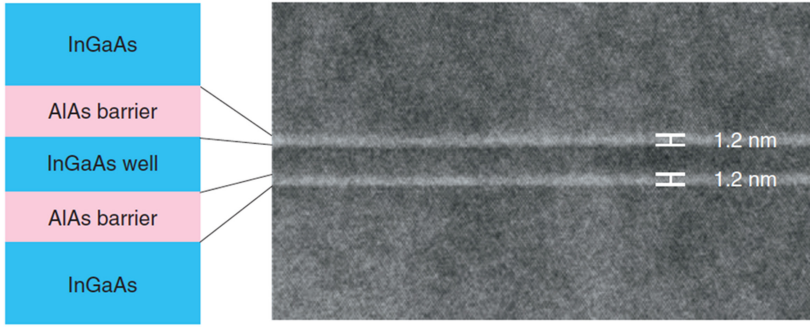


Figure 2.2. This shows the schematic of InGaAs/AlAs double barrier RTD and the cross-sectional transmission electron micrograph (TEM) image of the double barrier RTD. Adapted from [4].

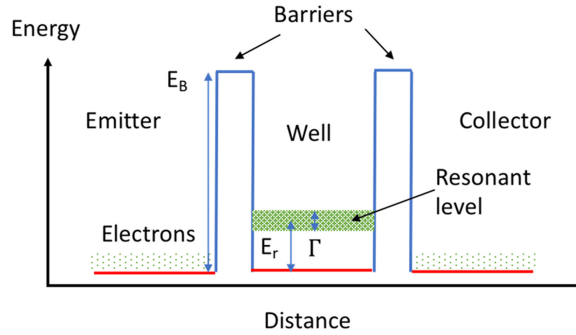


Figure 2.3. Schematic of the conduction band profile of the generic double barrier RTD with no voltage applied. The bottom of the conduction is plotted as a function of distance. The low bandgap material (red) for example InGaAs forms a quantum well between two barriers of high bandgap material (blue), for example AlAs, Γ is the width of the electronic bound state resonance at energy E_r .

As illustrated in figure 2.4, when a positive emitter voltage is applied to the RTD, the bottom of the conduction band in the emitter region is raised in energy relative to the collector side until it reaches the same energy as the resonance level, at the voltage level, V_p , this is the resonance condition and the RTD reaches the peak current flow, I_p . When the voltage is increased beyond, $V > V_p$, the RTD is in an off-resonance state and the current drops, reaching its lowest value I_v at the voltage V_v .

The physics which, in detail, explains the RTD current–voltage I – V , starts with a consideration of the various quantum mechanisms or channels open to the electrons as they make their way through the RTD structure. There has been a great deal of theoretical interest in the RTD and a good review can be found in [5]. Here we adopt a simplified explanation which is sufficient to describe the broad features of the RTDs analysed here, and with the help of some adjustable parameters, provides a reasonably good fit to the experimental I – V curve. It is based on references [2] and [6].

A general principle of quantum transport in mesoscopic systems is the Landauer formula [7], $G = G_0 \sum_i^{All} T_i(E, V)$, where G is the electrical conductance and G_0 is a

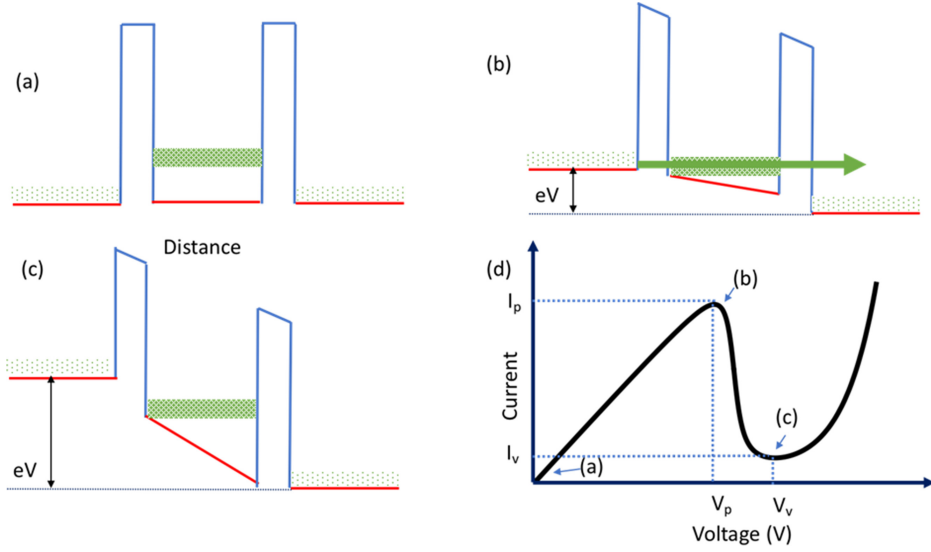


Figure 2.4. Conduction band profiles at three different bias voltages: (a) zero applied voltage $V = 0$, (b) resonance ($V = V_p$), (c) off-resonance ($V = V_v$), and (d) schematic of the current–voltage curve of the RTD showing how the curve relates to the conduction band profiles.

constant, and $T_i(E, V)$ a channel transmission coefficient. This formula relates the electrical conductance of a quantum conductor to the sum of the transmission coefficients for the various transmission channels. In what follows we divide the transmission channels into just two channels: a resonant channel and a nonresonant channel. The resonant channel accounts for transmission through the resonant state in the quantum well. All the other possible channels, for example, transmission by high energy electrons over the barriers are lumped into nonresonant channels. It turns out that this treatment is sufficient to account for the main features of the I – V curve feature and, with the aid of fitting parameters, can be adapted to fit the measured I – V curves.

For transport through the resonance level in the quantum well, the standard formula for RTD, the current density in the effective mass approximation is given by [2]:

$$J_R = \frac{|e|m^*}{2\pi^2\hbar^3} \int_0^\infty T(E, V) \ln \left[\frac{1 + e^{(E_F - E)/kT}}{1 + e^{(E_F - E - eV)/kT}} \right] dE \quad (2.1)$$

where $|e|$ absolute value of charge of electron, m^* is the effective mass on the electron in the semiconductor, $\hbar$ is Planck's constant/ 2π , E is the energy measured from the conduction band edge, E_F is the Fermi energy. $T(E, V)$ is given by:

$$T_r(E, V) = \frac{\left(\frac{\Gamma}{2}\right)^2}{\left[E - \left(E_r - \frac{eV}{2}\right)\right]^2 + \left(\frac{\Gamma}{2}\right)^2} \quad (2.2)$$

where Γ is the energy resonance width and E_r is the energy of the resonant level relative to the bottom of the quantum well.

Generally, the width of the quantum well resonance, Γ , is small compared to room-temperature thermal energies and so $T_r(E, V)$ is negligible except when the $E = \left(E_r - \frac{eV}{2}\right)$ applies. So, we can substitute $\left(E_r - \frac{eV}{2}\right)$ for E and take it outside the integral. The integral is then solved to get:

$$J_R(V) = \frac{|e|m^*T\Gamma}{4\pi^2\hbar^3} \ln \left[\frac{1 + e^{(E_F - E + eV)/kT}}{1 + e^{(E_F - E - eV)/kT}} \right] \cdot \left[\frac{\pi}{2} + \tan^{-1} \left(\frac{E_r - \frac{eV}{2}}{\frac{\Gamma}{2}} \right) \right] \quad (2.3)$$

So, this expression of the tunnelling current gives the right form of the equation that can be related back to the physics. To actually model a given I - V curve, adjustable parameters are introduced as follows:

$$J_r(V) = A \ln \left[\frac{1 + e^{(B - C + n_1 V)e/kT}}{1 + e^{(B - C - n_1 V)e/kT}} \right] \cdot \left[\frac{\pi}{2} + \tan^{-1} \left(\frac{C - n_1 V}{D} \right) \right] \quad (2.4)$$

where A is in A cm^{-2} , B , C and are in V and n_1 is dimensionless. There are other nonresonant channels that allow the electron to make its way through the double barrier RTD structure such thermal excitation over a barrier and these give rise to a valley current which be taken account of in the following expression:

$$J_{NR}(V) = H(e^{(n_2 V)e/kT} - 1) \quad (2.5)$$

where H is in A cm^{-2} and n_2 is dimensionless.

The total current density flowing through the RTD is given by $J(V) = J_R(V) + J_{NR}(V)$. Figure 2.1 shows the fit from [2] for a RTD made from a 4.8 nm well $\text{In}_{0.53}\text{Ga}_{0.47}\text{In}$ with 2.6 nm AlAs barriers. The values of the fitting parameters used were $A = 10^4 \text{ A cm}^{-2}$, $B = 0.035 \text{ V}$, $C = 0.1472 \text{ V}$, $D = 0.0052 \text{ V}$, $n_1 = 0.115$, $H = 0.1411 \text{ A cm}^{-2}$, $n_2 = 0.120$. This method allows any RTD I - V curve to be fitted but the fitting parameters obscure some of the underlying physics and only really hints at how the RTD design can be adjusted for any given application.

The final form of RTD current-voltage curve [8] is given by:

$$F(V) = I_R(V) + I_{NR}(V) = M[J_R(V) + J_{NR}(V)] \quad (2.6)$$

where M is a scaling factor that takes into account the area of RTD device.

For many applications a large peak I_p to valley current I_v ratio, $\left(\frac{I_p}{I_v}\right)$, is required and that can be obtained with large energy barriers height, E_B . If we compare figure 2.5(a) the I - V curve of the AlGaAs/GaAs RTD with figure 2.5(b) the I - V curve of the InGaAlAs RTD, it is apparent the InGaAlAs RTD has a larger $\left(\frac{I_p}{I_v}\right)$, which can be attributed to the larger energy barriers height, E_B , and the smaller electron effective mass of the InGaAlAs RTD.

In figure 2.5, there are regions in the I - V curve where the device oscillates. In the I - V curve the oscillation regions are where there are plateaus in the I - V curve at

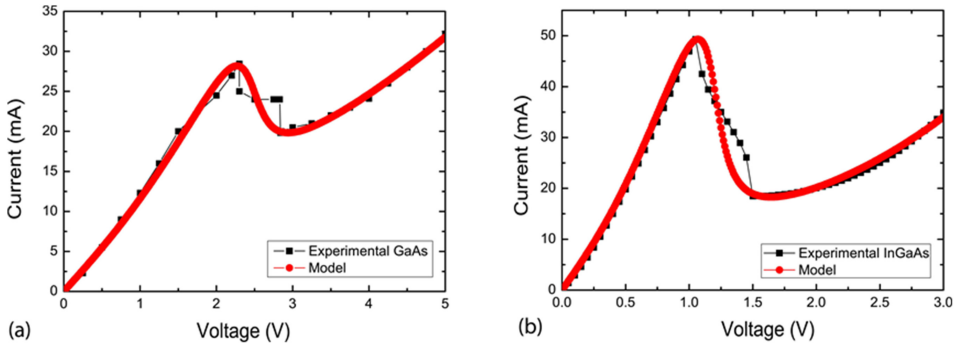


Figure 2.5. The current–voltage curves with model fit of (a) AlGaAs RTD and (b) InGaAlAs RTD. The red line fit is from model summarised in equation (2.4), for (a) $M = 2 \times 10^{-6} \text{ cm}^2$, $A = 45\,000 \text{ A cm}^{-2}$, $B = 0.05$, $C = 0.0874$, $D = 0.007$, $n_1 = 0.0350$, $H = 18500 \text{ A cm}^{-2}$, (b) $M = 1 \times 10^{-6} \text{ cm}^2$, $A = 200\,000 \text{ A cm}^{-2}$, $B = 0.068$, $C = 0.1035$, $D = 0.009$, $n_1 = 0.0350$, $H = 8500 \text{ A cm}^{-2}$.

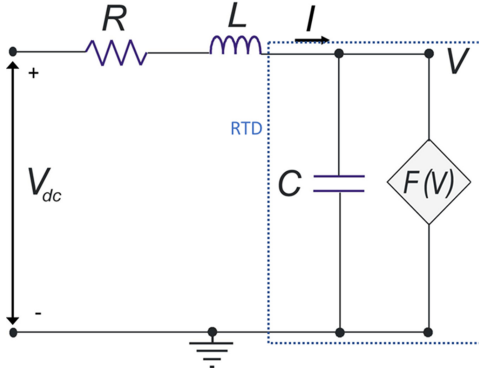


Figure 2.6. The DC bias circuit for the RTD, the R and L lumped components represent the parasitic components associated with the leads supplying the DC voltage to the RTD—the equivalent circuit for the RTD (inside the dashed blue box) is a capacitance in parallel with a voltage dependent current source $F(V)$.

voltages great than V_p . In this region of the I – V curve the RTD has electronic gain also the parasitic circuit elements, associated with the direct current (DC) bias circuit, provide feedback. The RTD has a large bandwidth (see discussion in next section) and therefore there is often a frequency with sufficient feedback and gain to cause the RTD to oscillate. The methods of obtaining an I – V curve of a tunnel diode under stable non-oscillatory conditions are discussed in [9]. These methods are, in essence, choosing the correct value of a series resistance in the RTD DC bias circuit.

The corresponding alternating current (AC) circuit of an RTD oscillatory circuit is depicted in figure 2.6. The time differential of the voltage $\dot{V}$ is given by:

$$\dot{V} = \frac{1}{C}[I - F(V)] \quad (2.7)$$

and the current is given by

$$\dot{I} = \frac{1}{L}[V_{\text{DC}} - RI - V] \quad (2.8)$$

After some algebraic manipulation of the above equations, the following second-order differential equation can be obtained:

$$\ddot{V} + H(V)\dot{V} + G(V) = 0 \quad (2.9)$$

with $H(V) = \frac{R}{L} + \frac{1}{C} \frac{dF(V)}{dV}$ and $G(V) = V - V_{\text{DC}} + RF(V)$. Equation (2.9) is a nonlinear second-order differential equation which can be used to describe a Liénard system [10]. The Liénard system is a more general case of the more familiar Van der Pol oscillator [11] system of equations (2.7)–(2.9). In the Van der Pol case the term multiplying the first-order differential, $\dot{V}$, is a third-order polynomial and with the Liénard system this nonlinear term can be more general, for example, a higher order polynomial. If we evoke the analogy of a mechanical oscillator, then $G(V)$ is a nonlinear force and $H(V)\dot{V}$ is a damping factor.

When biased to the NDC region—the circuit illustrated in figure 2.6 acts as a relaxation oscillator operating at a fundamental frequency close to $f_0 \sim 2\pi(LC)^{-\frac{1}{2}}$, the condition for operation is that the series resistance is less than the absolute value of the negative resistance $R < |NDR|$.

2.2.1 High-frequency response

A key driver of research interest in resonant tunnelling devices is the potentially ultrahigh-frequency response of the device, and fundamentally, the high-frequency response [3] can be traced back to the tunnelling time because the maximum frequency response cannot exceed the inverse of the tunnelling time. The tunnelling time sets the upper speed limit [5]. Theoretically estimations of the tunnelling time through an RTD structure are in the range 10–100 fs [12] and so, because the tunnelling time is so short, in any practical RTD circuit the real speed limit is *de facto* set by the parasitic components associated with the bias and feedback circuit.

Experimental investigation of the switching speed is consistent with picosecond time scale for RTD switching [13] and it has proved possible to determine separately the contribution of parasitic or resistance/capacitance (RC) contribution and the more fundamental contribution to the switching time from the quantum well quasibound state lifetime [14].

A simple, intuitive approach to estimate the tunnelling time is to estimate the time the tunnelling electron takes to cross the width of the barrier. This is calculated from the velocity of the electron as determined from its k vector and the width of the barrier. Thus, the tunnel time is given as follows [15]:

$$t_{\text{tun}} = \left(\frac{2E}{m^*} \right)^{-1/2} W \quad (2.10)$$

where W is the barrier width and E is the electron energy. This was one of the original approaches to estimate the tunnel time through a single barrier.

To illustrate the above considerations we take as an example of a high-frequency design of an RTD device with 1.9 THz oscillation frequency and $\sim 0.4 \mu\text{W}$ output power [3]. The design principles can be summarised as:

- reduce the thickness of the RTD layers: the AlAs barriers are 1 nm thick and the InGaAs well is 2.5 nm;
- reduce the parasitic component values: so the inductive and resistive load of the slot antenna for the THz radiation is minimised by using a micro-fabricated slot antenna and using highly doped contact layers (InGaAs n-type $\sim 5 \times 10^{19} \text{ cm}^{-3}$) 9 nm thick;
- for high power output at high-frequency—a large gain bandwidth—then a large peak to valley ratio is required and the high AlAs barriers ensure a deep quantum well which reduces the valley current by reducing thermal energy induced leakage of charge carriers over the barriers.

2.3 Optical properties: optical waveguiding, electroabsorption, photoconduction and lasers

The optical properties of the RTD photonic devices are dominated by the optical properties of the epitaxial semiconductor alloy layers that make up the device. The actual RTD structure itself consists of layers which are much thinner than the wavelength of light in the semiconductor. For example, a typical RTD structure including barriers and quantum well is <10 nm thick and RTD photonic devices designed for optical telecommunications wavelengths (~ 1500 nm) have an optical wavelength of around 400 nm in the device. So, the RTD structure is only a minor perturbation on the optical properties. The RTD is mostly included to provide very high-speed switching and/or high bandwidth NDR, as thoroughly discussed in the electrical properties section 2.2. The RTD photonic devices have been fabricated with optical functionalities that include optical waveguide, electroabsorption, photoconduction and lasers—the device optical functionality is provided by the other semiconductor alloy epilayers which make up the device.

For example, if an RTD electroabsorption optical modulator is required then the device needs semiconductor epilayers that provide optical waveguiding so that the light is confined to a core layer for the optimum interaction length and the core layer has its bandgap energy tuned to the value appropriate for maximum electro-absorption effect.

In this chapter, we give a general description of the optical properties that are useful for a RTD photonics device. In subsequent chapters, more specific descriptions are given of how exactly these properties are employed in the following devices:

- (i) the optical waveguide RTD electroabsorption modulator (RTD—EAM),
- (ii) the RTD photodetector (RTD—PD),
- (iii) the RTD laser diode (RTD—LD).

2.3.1 Optical waveguiding

Semiconductor electroabsorption modulators [16] can be employed in optical communication systems. Typically they are optical waveguide devices because in

waveguide devices the interaction length between the core semiconductor epilayer and the light is not limited by diffraction as would be the case in bulk optics. Diffraction limits the interaction region in bulk optics because the focal region is inversely proportional to the width of the focus so a tightly focused beam rapidly expands. In a waveguide configuration the focal region is captured by the waveguide and the tight focus is maintained for the length of the waveguide. The waveguide device is limited not by diffraction but losses in the waveguide, be they scattering or absorption losses. The RTD—EAM described in chapter 3 is a waveguide and here we describe the basic principles of optical waveguiding for semiconductor photonic devices.

Figure 2.7 shows a schematic of a generic semiconductor optical waveguide. The epilayers are semiconductor alloys; for example, for an AlGaAs based device, typical parameters (for a 1550 nm wavelength operation) are as follows; the core layer refractive index $n_c \approx 3.5$ and thickness $d_c \approx 1550$ nm; upper cladding refractive index $n_{uc} \approx 2.9$ and thickness $d_{uc} \approx 2000$ nm lower cladding refractive index $n_{lc} \approx 3.30$ and thickness $d_{lc} \approx 3000$ nm; the height of the channel $h \approx 300$ nm and width $w \approx 3750$ nm. A simple effective index method (see, for example, [17]) can be used to determine the waveguide modes that determine how light is confined in the waveguide.

The effective index method gives a first approximation to the type of optical mode that will propagate in a semiconductor optical waveguide. The result of an effective index method calculation for an AlGaAs waveguide is shown in [18]. A much more detailed analysis that takes into account the full vectorial nature of the light is found in [19]—this paper has an associated website from which software can be downloaded [20] to calculate the waveguide modes.

The key advantage conferred by the waveguide configuration is that the interaction length between light and the semiconductor material in the core is not limited by diffraction but by the optical losses which can be due to waveguide

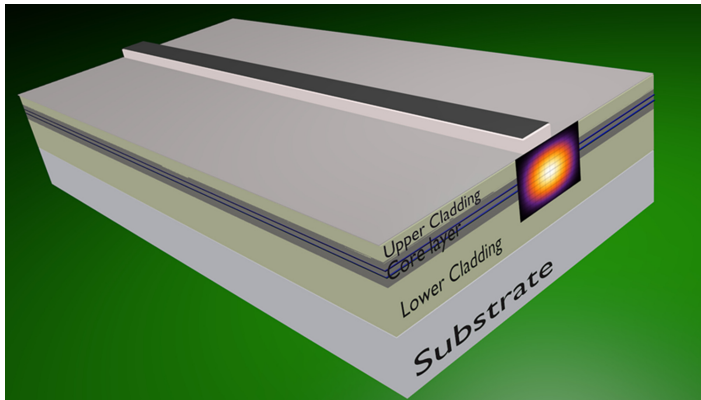


Figure 2.7. Layout and epilayer structure of a generic RTD semiconductor channel waveguide. At the end of the waveguide the approximate distribution of optical intensity for a single mode waveguide is shown. The thin blue layers indicate the barrier layers of the RTD in the centre of the core.

scattering from, for example, rough walls on the channel mesa or from intrinsic losses from absorption by the semiconductor. The increased interaction length is important for electroabsorption devices and can be important in photodetector devices. At photon energies just below the bandgap energy of the semiconductor, dominated by scattering losses, the semiconductor waveguide loss has been measured to be around 20 cm^{-1} [21].

Another advantage of the waveguide configuration is the compact layout of the electrodes in the dimension perpendicular to the plane of the layers. So that, for example, a small applied voltage of a few volts can result in electric fields of a few kV per cm in the core of the waveguide useful for both electroabsorption and electro-optic devices.

2.3.2 Electroabsorption

The RTD can be used to efficiently switch the electric field in the core part of a semiconductor optical waveguide—the exact mechanism is discussed in chapter 3. The change in electric field can be used to alter the absorption of light in the core of the semiconductor optical waveguide and thereby encode the light with information. The nonlinear electrical switching characteristics of the RTD, as illustrated in figure 2.1, means that the RTD electric field switching is best for digital modulation rather than analogue optical modulation.

The Franz–Keldysh effect refers to electrical field dependent change in the optical absorption close to the band edge of a bulk semiconductor. From a consideration [18, 22, 23] of how an applied electrical field induces overlap between the electron valence band and conduction band wavefunctions in a material, the following expression for the absorption coefficient spectrum can be derived

$$\alpha(\omega, F) = \left(\frac{C_p \theta_F^{\frac{1}{2}}}{\omega} \right) [|\acute{A}i(\eta)|^2 - \eta |Ai(\eta)|^2] \quad (2.11)$$

where $\theta_F^3 = \frac{e^2 F^2}{2m_{vc}\hbar}$, $\eta = (E_g - \hbar\omega)/\hbar\theta_F$, F is the applied electric field, C_p is a material constant which is similar for III–V semiconductors and CGS units is given by $C \sim 1.76 \times 10^{12} \text{ cm}^{-1} \text{ s}^{-1/2}$ —actually, this value is for GaAs. The parameter m_{vc} is the reduced mass of the electron–hole pair for GaAs, $m_{vc} = 0.065 m_o$, where m_o is the mass of the free electron; ω is the optical angular frequency, E_g is the bandgap energy; $Ai(\eta)$ is the Airy function, $\acute{A}i(\eta)$ is the differential of the Airy function (Airy prime function).

The optical spectrum calculated from equation (2.11) is shown in figure 2.8 for the alloy $\text{In}_{1-x-y}\text{Ga}_x\text{Al}_y\text{As}$ with the $y = 0.05$ the fraction set so that the bandgap energy is close optical telecommunications ‘C band’ photon energies (wavelengths around 1550 nm) [24].

It turns out that the Franz–Keldysh effect of bulk semiconductors is not the optimum electroabsorption effect for electroabsorption modulators. It is better to employ multiple quantum wells in a layered stack in which case the electro-absorption effect is termed the quantum confined stark effect (QCSE) [25]. When the

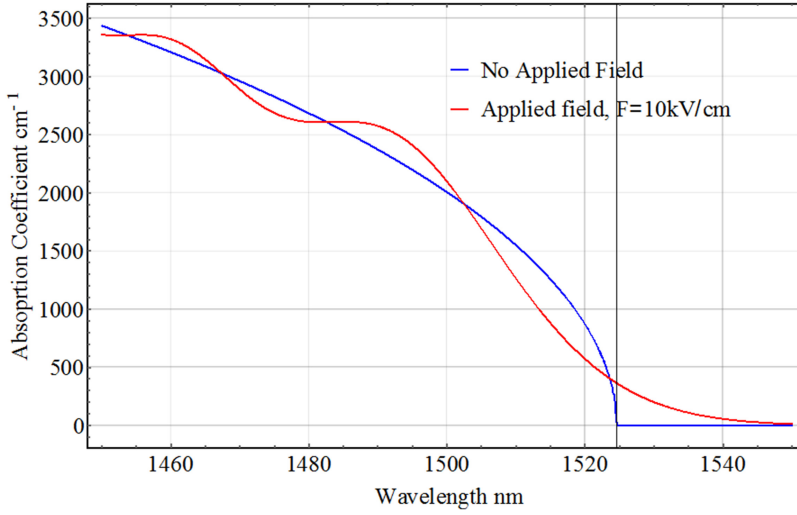


Figure 2.8. Franz–Keldysh effect in InAlGaAs. Calculated optical absorption spectrum of a semiconductor alloy $\text{In}_{1-x}\text{Ga}_x\text{Al}_y\text{As}$ [24] $y = 0.05$ with and without an applied electric field.

quantum wells are optimised to enhance the QCSE for a given wavelength then the best performing EAMs are produced [25, 26].

2.3.3 Photoconductivity and photodetection

The I – V curve of the RTD is, to some extent, dependent on the absorption of light in the device because the conductivity of the epilayers is increased by the creation of electron–hole pairs as in the photoconductivity effect [27]. RTD devices can be designed to utilise photoconductivity to produce highly sensitive photodetectors [28] in the visible and near infrared. Recently the concept has been extended to include the mid-infrared [29]. We give a simple analysis on how a double barrier RT structure can be used to amplify the current in a photoconducting device.

Photons with energy greater than the bandgap energy create mobile charge carriers in a layer of the RTD device (note: much lower carrier numbers are generated in the RTD barriers and quantum well which are generally too thin to absorb a significant amount of light) and thereby increase the electrical conductivity of the photoabsorption layer. A simple way to analyse [30] is to include a light dependent resistance component into the circuit as shown in figure 2.9. The total resistance of the RTD device internal resistance now becomes:

$$R_s = R_i + R(p) \quad (2.12)$$

where the total resistance has been split into a light dependent part, $R(p)$, and a part independent of the incident light R_i .

The voltage, V' , across the double barrier structure is given by:

$$V' = V - R(p)_s I \quad (2.13)$$

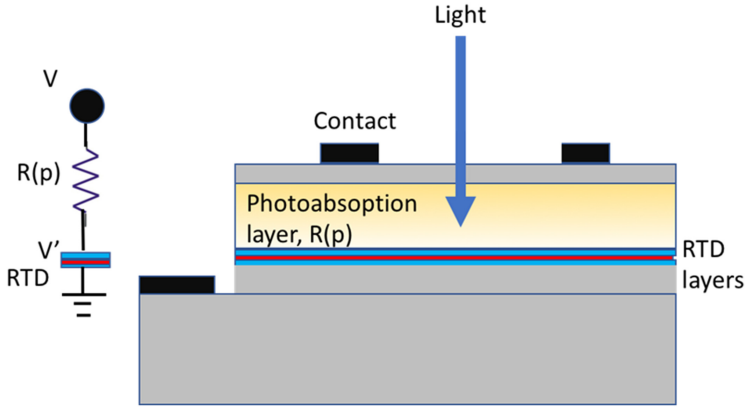


Figure 2.9. The lumped element electrical circuit and the layer structure of a generic resonant tunnelling diode photoconductive circuit. A voltage, V , is applied to bias the circuit and voltage, V' , which depends on the light intensity modulated resistance, $R(p)$, appears at the double barrier RTD structure.

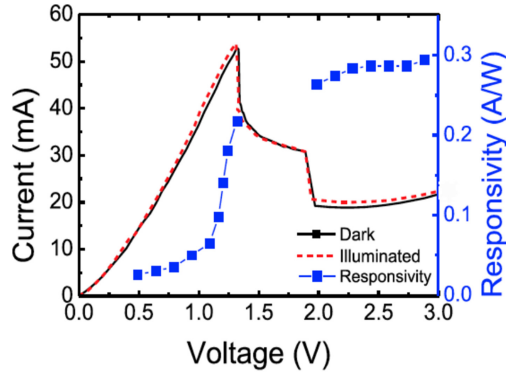


Figure 2.10. The dark and illuminated I - V curve for a waveguide RTD—PD with a InGaAlAs photo-absorption layer and a AlAs/InAlAs/AlAs RTD are shown. The effective responsivity versus applied bias voltage for light with a wavelength of 1550 nm. is also shown. Adapted from [28].

where I is the current in the device.

When light is incident on the device the effective voltage, V' , at the double barrier is increased as the resistance, $R(p)_s$, is reduced by photoconductivity. If the device is biased close to resonance then, with a small change in incident light, the device can jump from on- to off-resonance, that is, from V_p to V_v (see figure 2.4), thus providing amplification. Therefore, the effective responsivity of the RTD photodetector (RTD—PD) is increased when the device is biased close to V_p . This is illustrated in figure 2.10 where it is shown that the responsivity as a function of the applied voltage can be enhanced by a factor ~ 10 when the device is biased in the valley region.

Subsequently, more refinements [31] to the RTD—PD have achieved responsivities as large as $3.1 \times 10^4 \text{ A W}^{-1}$ and the ability to detect a few (~ 26) photons. The responsivity is large enough that the RTD—PD can be used to directly drive other

optoelectronic devices such as laser diodes [32] and thus build systems such as the optoelectronic oscillator [33].

There are other performance metrics apart from responsivity and these will be covered in more detail in chapter 4 on the RTD—PD.

2.3.4 Optical emission

The RTD can also be incorporated into semiconductor devices that emitted light. RTDs have been incorporated into laser diodes (RTD—LD) [32] and light emitting diodes (LEDs)—RTD—LED [34]. Usually the light emission requires that the RTD is combined with a *pn* junction but it has also recently been observed in devices without *p* type material and operates via impact ionisation [35].

The RTD confers a NDC region to the electrical characteristics of the RTD—LD and RTD—LED. The NDC region can be used to amplify an applied modulation signal so that the optical output from the device can be modulated with a low power, higher frequency, signal. For the RTD—LED this can lead to a modulation figure of merit, M (in $\mu\text{W}/\text{mA} \times \text{MHz}$) increasing from $M \sim 6$ in heavily doped LEDs to $M \sim 4400$ in RTD—LEDs [36].

In chapter 5, we have an extensive discussion on how the RTD—LD improves the modulation of laser diodes.

2.4 Conclusions

This chapter has been an introduction to the essential electrical and optical characteristics of semiconductor optoelectronic devices that include a double barrier quantum well RTD structure. The unifying theme is that the RTD provides very high bandwidth electrical gain in a very compact epilayer structure (~ 10 nm) that is easy to incorporate into another semiconductor epilayers. This enables unique photonic device functionalities in photodiodes, lasers and modulators that can be driven with low amplitude and high-speed input signals taking advantage of the wide bandwidth and on-chip electrical gain of RTD integrated devices.

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Resonant Tunneling Diode Photonics

Devices and applications

Charlie Ironside, Bruno Romeira and José Figueiredo

Chapter 3

Resonant tunnelling diode—electro-absorption modulator (RTD—EAM)

Optical communication systems relying on microwave/millimetre-wave-optical transducers are fundamental to the development of applications such as low-cost fibre-optic communication networks, radio-over-fibre wireless communications, and radio phased array antennas. Those systems require the use of very fast optical modulators [1] since laser diodes cannot be directly modulated at frequencies beyond their resonance frequency which is typical below few tens of gigahertz [2]. In this chapter, we cover our work on low driving voltage high-speed and highly-efficient electrical active electro-absorption modulators (EAMs). The EAM is based on the integration of a double barrier quantum well (DBQW) structure within AlGaAs/GaAs and InGaAlAs/InGaAs unipolar semiconductor optical waveguides and operates at the optical windows corresponding to wavelengths around 900 nm, 1300 and 1550 nm. This novel device is known as a resonant tunnelling diode electro-absorption modulator (RTD—EAM).

3.1 Introduction

Optical communication systems capable of supporting very high data bit rates per channel comprise a lightwave carrier source, typically a laser diode, a transmission media such as the air or an optical fibre (eventually utilizing in-line optical amplification), and a lightwave receiver incorporating a photodetector such as a photodiode (PD) or an avalanche photodiode (APD). The process that allows a lightwave to convey information is known as optical modulation and corresponds to the intentionally change of at least one property of the lightwave according to the electrical signal that carries the data to be transmitted. A lightwave carrier can be modulated in amplitude (or intensity), frequency, polarization, and phase, with the choice depending upon the application and on the technologies available.

Assuming that the bandwidth of the optical communication system is not limited by the receiver capacity to convert the intentional changes of the lightwave property, such as the intensity, into an electrical signal, the higher the speed the optical modulation can be performed the higher is the data bit rate the system is capable of handling.

The direct modulation of a laser diode using a driving current cannot achieve data bit rates much higher than few tens of gigabit per second (Gbit s^{-1}). Typically, direct modulation is used to realise data bit rates well below 20 Gbit s^{-1} [2], because direct modulation of laser diodes beyond a few Gbit s^{-1} leads to several undesirable effects such as optical frequency chirping (due to refractive index modulation induced by the current used to modulate the laser diode). The chirp parameter is defined as the ratio of the changes in the real and imaginary part of the refractive index caused by the laser injection current. Therefore, it causes linewidth broadening, which introduces distortion in the transmitted signal, limiting system performance, particularly in long distance communications and at very high data rates. To attain data bit rates beyond a few tens of Gbit s^{-1} external modulation schemes need to be considered.

3.2 Electro-optic and electro-absorption modulators

External modulation schemes rely on a continuous lightwave (a lightwave produced by the unperturbed laser diode) travelling through a material medium which can be intentionally modified by externally inducing changes in the complex refractive index of the material. The external agent can be, for example, an electrical field. If a fast-changing electric field can induce changes at the same rate in the real and/or imaginary parts of the material complex refractive index, these fast changes lead to changes in at least one lightwave measurable property such as the phase, intensity, polarization or direction of propagation.

External modulators can be implemented either in waveguide configuration or bulk in transverse transmission or reflection configurations [3]. The waveguide types are the most appropriate for compact transmitter and photonic integrated circuit implementations, in which waveguides are used to guide the input and output lightwaves between devices along the surface of the substrate.

3.2.1 Electro-refraction modulators

When the relevant changes induced by the electric field are in the real part of the refractive index, the effect is known as electro-refraction (or linear electro-optic) and leads essentially to lightwave phase changes, that is, the lightwave induced phase change is directly proportional to the magnitude of the applied voltage, V : $\Delta\phi(V) \propto \Delta n \times V$. The linear electro-optic effect is used to implement phase modulators or intensity modulators, e.g. in an interferometer configuration, as represented in figure 3.1(a) [4, 5].

In electro-refraction modulators based in the linear electro-optic effect an applied voltage induces a proportional change of the material refractive index which leads to a variation of the lightwave phase. This behaviour is used to implement

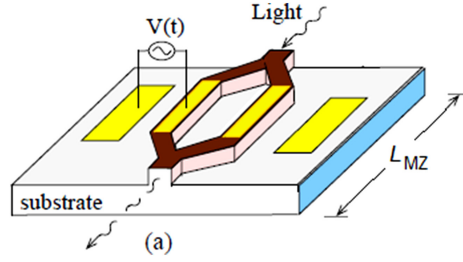


Figure 3.1. Typical external intensity modulator in a waveguide Mach–Zehnder interferometric configuration based on linear electro-optic (electro-refraction) effect.

interferometric structures for intensity modulated lightwaves. The most common electro-refraction modulators have a Mach–Zehnder configuration, figure 3.1, implemented using ferroelectric materials such as lithium niobate (LiNbO_3) because they have strong electro-optic effect and are dielectric. This leads to low loss and marginally associated heating effects, when compared with semiconductor materials. However, modulators based on ferroelectric materials are several centimetres long because the need for interferometer or directional coupler configurations. Moreover, they require driving voltage amplitudes in excess of 5 V at very high frequency. As a consequence, monolithic integration with photonic integrated circuits is challenging. Despite the limitations for compact solutions, bandwidths in excess of 100 GHz can be achieved employing traveling-wave designs in which the electrical signal is propagated on a micro-strip line along with the lightwave.

3.2.2 Electro-absorption modulators (EAMs)

When the electric field changes, this leads to changes in the imaginary part of the refractive index, related to the material absorption coefficient, α . The effect is known as electro-absorption and leads to the modulation of the transmission coefficient of the material. This effect enables the modulation of intensity of the lightwave, that is, the material transmission coefficient, T , which is function of the absorption coefficient, α , is modulated by a changing applied voltage V : $T(V) \propto \exp[-\alpha(V) \cdot L]$, where L is the thickness of the portion of the material in the direction of the light propagation. Figure 3.2 shows schematically a waveguide electro-absorption external modulator.

The electro-absorption modulators (EAMs) considered here are implemented using direct bandgap semiconductor materials. Typically, they are much smaller than interferometric electro-refraction modulators. Further, because of their smaller size, compatibility with CMOS technologies and response times in the sub-picosecond, electro-absorption modulators based on semiconductor waveguides, figure 3.2, are interesting components for on-chip high-bandwidth optical communication links particularly when integrated with laser diodes. Assuming the parasitics of these components can be substantially reduced, the bandwidth of electro-absorption modulators can reach hundreds of GHz.

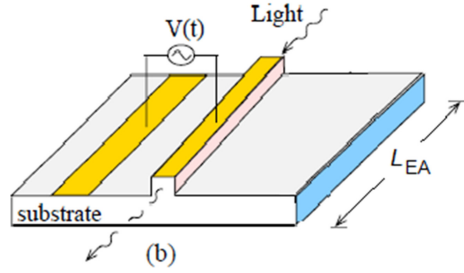


Figure 3.2. Typical configuration of a intensity modulator employing the electro-absorption effects (either Franz–Keldysh or quantum confined Stark effects). Typically, $L_{MZ} \gg L_{EA}$.

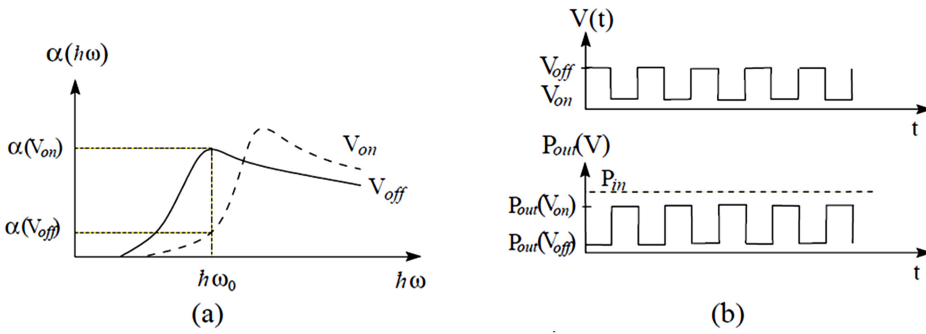


Figure 3.3. Schematic representation of the operation of EAM: (a) absorption coefficient at two voltage values, showing the band-edge broadening due to the Franz–Keldysh or quantum confined Stark effects; (b) the applied voltage and the modulator output power. $\hbar\omega$ is the incident light photon energy; V_{on} and V_{off} represent to the voltage values corresponding to the on and off states.

Waveguide electro-absorption modulators are designed to operate at optical wavelengths slightly below the bandgap of the low doped waveguide core, or intrinsic material, making them transparent at these wavelengths. Under an applied voltage a band-edge shift is observed, resulting in an increase of the lightwave propagation loss, e.g. the applied voltage leads to an intensity change of the waveguide output. Figure 3.3 shows a diagram describing the principle of operation of an electro-absorption modulator.

Although potentially with a small footprint, compatible with CMOS technologies and enabling a response time in the sub-picosecond range, electro-absorption modulators still require a large voltage. The applied voltage needed for an on/off ratio of about 10 dB is typically larger than 2 V (peak-to-peak). A limiting factor in the high-speed response is the time constant for the holes to be swept out of the waveguide core.

3.3 Resonant tunnelling diode EAM device

As discussed previously, when the RTD operation point switches from the peak to the valley region a strong electrical field enhancement is achieved as a result of the current decrease from the peak to the valley. Assuming there is no change in the

series resistance, this electrical field strengthening results from the internal voltage drop across the depletion layer as a consequence of the current reduction through the NDC.

The electric field distribution dependence on the applied voltage in the direction of the carrier flow in a RTD structure can be understood considering the Γ -conduction band profile of the injection and collector spacer layers as schematically shown in figure 3.4. Below resonance (first positive differential conductance (PDC) region), the applied voltage is dropped mainly across the DBQW and the adjacent regions particularly on the collector side, figure 3.4(a), and the electric field varies accordingly. Above resonance (in the NDC and in the second PDC region), the additional applied bias voltage enhances the depletion of the collector region reinforced also by the current drop, which leads to a larger depletion region, figure 3.4(a). In addition, there is an electric field enhancement. Most of the additional applied voltage beyond the peak is dropped across the depletion region formed between the second barrier and the collector contact, figure 3.4(b), where a strong electric field builds-in and/or extends across the collector low doped region.

External inducing peak-to-valley transitions or biasing in the NDC an RTD inserted in a resonant tank circuit leads to the modulation of the interband absorption coefficient for light with photon energies close to but smaller than the collector bandgap energy through the Franz–Keldysh effect [6, 7]. However, in typical RTD implementations this effect leads to almost negligible modulation depths because the thickness of the depleted region is at least one order of magnitude lower than the operating wavelength. Such a small light path means that a transverse transmission electro-absorption modulator would produce an extremely small absorption enhancement.

In the transverse transmission configuration, the light is injected perpendicularly to the tunnelling plane that makes the light absorption volume quite small. Furthermore, the change of the electric field induced by the peak-to-valley switching leads to a very small modulation of the absorption coefficient, and thus, a small modulation depth is expected. This limitation is easily overcome by increasing the

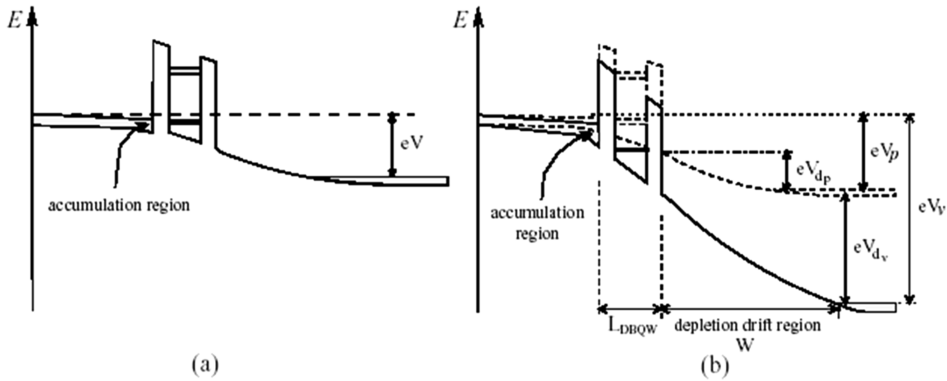


Figure 3.4. Effect of applied biased on RTD Γ -band: (a) before the peak and (b) on the valley.

thickness of the undoped and low doped regions adjacent to the DBQW in a waveguide configuration where light propagates in the plane of the DBQW. The light absorption volume can be substantially increased by increasing the light path in the direction parallel to the barrier-depletion region interface plane. This new modulator structure is called resonant tunnelling diode electro-absorption modulator (RTD—EAM). The RTD—EAM requires much lower driving voltage for a given modulation depth by taking advantage of the electrical gain provided by the N-shape current–voltage characteristic.

The combination of the fast switching capabilities with a broadband electrical gain of RTD structures enables the implementation of high-speed, highly-efficient and low-cost electro-absorption modulators. RTD—EAMs based on AlGaAs/GaAs for operation at optical wavelengths around 900 nm, and RTD—EAMs made of InGaAlAs/InP operating at 1550 nm were demonstrated as highly-efficient very low driving voltage electro-absorption modulators. The modulation efficiency characterized by the bandwidth-to-drive-voltage ratio, for a 10 dB modulation depth, was higher than 40 GHz V⁻¹ [8, 9]. The InGaAlAs/InP structures can be designed to operate in the wavelength range from 1200 nm to 1600 nm, depending of the aluminium concentration of the collector region adjacent to the barrier.

3.4 RTD—EAM operation principle

A waveguide electro-absorption modulator requires light confinement and guiding along the region where there is the largest electrical field enhancement. In the case of an RTD—EAM the DBQW structure is embedded in the core of the waveguide. The waveguide is sandwiched between two layers of a larger bandgap (lower refractive index) material forming the cladding layers. This new structure is identical to an RTD with two large spacer layers each side of the DBQW. Lastly, highly doped layers of a lower bandgap material are included (typical of the same material as the spacer layer) on each side of the epi-layer for ohmic contact purposes.

This unipolar semiconductor optical waveguide structure incorporating a DBQW is schematically shown in figure 3.5(a). The corresponding epi-layer Γ -conduction band-edge and refractive index profiles are also shown in figure 3.5(a). The device is called a resonant tunnelling diode optical waveguide (RTD—OW). As shown schematically in figure 3.5(b), this ridged waveguide channel configuration, confines the butt-coupled light along the directions that contains the tunnelling layers due to the refractive index distribution, therefore allowing a longer light interaction length (up to 100-fold larger than the light wavelength) with the region of the largest electric field.

The RTD—OW operation as an EAM can be easily understood if we consider the expected current–voltage characteristic of a similar unipolar waveguide structure where the DBQW structure is not included, i.e. leading to a device without the negative differential conductance region [10]. The DBQW included in the waveguide core strongly modifies the waveguide I – V leading to an N-shape current–voltage characteristic that is strongly nonlinear. As a result, a nonlinear electric field distribution across the waveguide core is achieved. The electric field is strongly

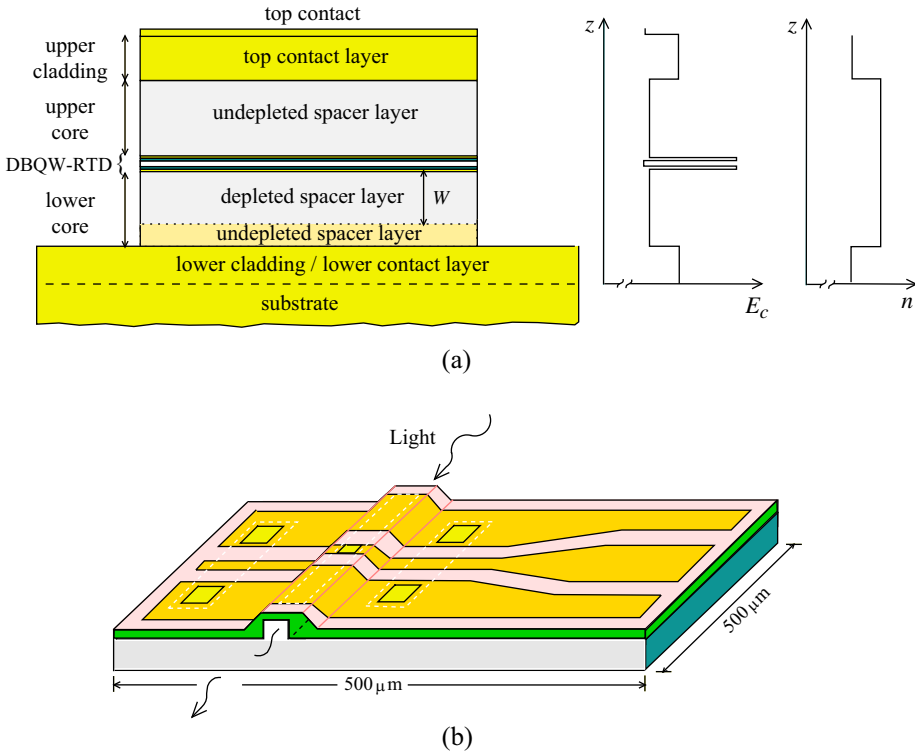


Figure 3.5. (a) Diagram of a unipolar RTD—OW wafer structure, and the corresponding Γ -conduction band-edge and refractive index profiles. (b) Ridged waveguide channel configuration. Such RTD—OW, apart from the light confining layers (the lower refractive index regions—top and bottom cladding layers), corresponds to a DBQW—RTD with very thick undoped/low doped emitter and collector spacer layers.

dependent on the bias voltage due to the electron accumulation close to the emitter barrier and due to the creation of a depletion region on the spacer layer of the collector side. When the device is biased in the valley region, light with energy slightly below the bandgap of the waveguide core is more highly absorbed. Figure 3.6 schematically shows the light absorption in an RTD—OW biased around the valley point. This is a result of the Franz–Keldysh effect in the depleted region of the collector side.

Since a small increase in the bias voltage can switch the RTD—OW operation point from the peak to the valley region, a large increase in the electric field across the collector depletion region can be achieved, resulting, through the Franz–Keldysh effect, in a pronounced absorption band-edge broadening effect. This leads to a substantial increase of the propagation losses in the waveguide at photon energies close to but smaller than the bandgap energy of the waveguide core material. This unique property can be exploited to implement either electro-absorption (EAM) (intensity) modulators or waveguide photo-detectors (as discussed in chapter 4).

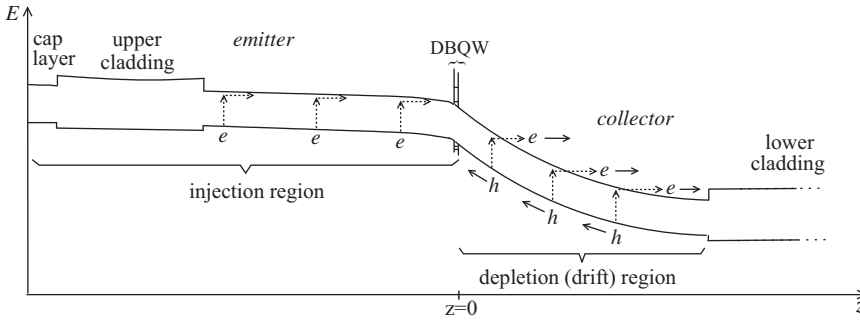


Figure 3.6. Schematic diagram of the bottom of the conduction band versus the top of the valence band versus distance illustrating the change in optical absorption induced by Franz-Keldysh effect in a RTD—OW biased around the valley point.

In order to estimate the absorption change we need to estimate the electric field enhancement induced by the peak-to-valley switching. For a given current density J , the magnitude of the electric field across the device depletion region as a function of the position, z , in the direction of the current, $F(z)$, can be obtained from the Poisson's equation, as:

$$F(z) = F_{0,J} - \frac{e}{\epsilon}(N_d^+ - J)z \quad (3.1)$$

where $F_{0,J}$ represents the electric field at the interface between the collector barrier and the collector spacer layer (where $z = 0$). The voltage drop across the collector region, ΔV_{PV} , due to the peak-to-valley switching is then given by:

$$\Delta V_{PV} = F_{0,J}W_{dep} + \frac{W_{dep}^2}{2\epsilon v_{sat}}\Delta J_{PV} \quad (3.2)$$

where $\Delta F_{0,J}$ represents the electric field enhancement due to the peak-to-valley switching at the interface between the collector barrier and the collector spacer layer; v_{sat} is the carrier saturation velocity; it is assumed that the thickness of the depletion (or drift) region, W_{dep} , does not change. From the relation (equation (3.2)) we can estimate the electric field enhancement, ΔF_{PV} , as a function of the peak-to-valley voltage difference as [9, 10]:

$$\Delta F_{PV} \cong \frac{\Delta V_{PV}}{W_{dep}} + \frac{W_{dep}}{2\epsilon v_{sat}}\Delta J_{PV} \quad (3.3)$$

The Franz-Keldysh absorption coefficient change as a result of the electric field enhancement due to the peak-to-valley switching is given by:

$$\Delta\alpha\left(\frac{hc}{\lambda}, \Delta F_{PV}\right)_{PV} = \alpha\left(\frac{hc}{\lambda}, F_V\right)_V - \alpha\left(\frac{hc}{\lambda}, F_P\right)_P \quad (3.4)$$

where $\alpha\left(\frac{hc}{\lambda}, F\right)$ is given, in the weak field approximation, by the electro-absorption coefficient resultant from the Franz–Keldysh effect—see section 2.3.2.

At a given photon energy, $\frac{hc}{\lambda}$, the absorption change induced by the electric field enhancement due to the peak-to-valley switching can be estimated as:

$$\Delta\alpha\left(\frac{hc}{\lambda}, \Delta F_{PV}\right)_{PV} \cong \alpha\left(\frac{hc}{\lambda}, \Delta F_{PV}\right)_V - \alpha_0\left(\frac{hc}{\lambda}\right) \quad (3.5)$$

where $\alpha_0\left(\frac{hc}{\lambda}\right)$ corresponds to the absorption loss of the material at wavelength λ . For simplicity of analysis, we assume that $\alpha\left(\frac{hc}{\lambda}, F_V\right)_V \gg \alpha\left(\frac{hc}{\lambda}, F_P\right)_P \cong \alpha_0\left(\frac{hc}{\lambda}\right)$, since $F_V \cong \Delta F_{PV} \gg F_P$.

The Franz–Keldysh absorption band-edge shift resultant from the peak-to-valley switching can be estimated using the relation:

$$\Delta\lambda_g \cong \frac{\lambda_g^2}{hc} \left(\frac{eh}{8\pi^2 m_r} \right)^{\frac{2}{3}} \Delta F_{PV}^{\frac{2}{3}} = \frac{\lambda_g^2}{hc} \left(\frac{eh}{8\pi^2 m_r} \right)^{\frac{2}{3}} \left(\frac{\Delta V_{PV}}{W_{dep}} + \frac{W_{dep}}{2e v_{sat}} \Delta J_{PV} \right)^{\frac{2}{3}} \quad (3.6)$$

The measurement of $\Delta\lambda_g$ enables evaluation the electric field change, ΔF_{PV} .

It is possible to design RTD—OWs that have a sharp current drop across the NDC and achieve a peak-to-valley switching with small applied voltages (<1 V) with high-frequency ac signals ($\gg 10$ GHz). In this case the voltage drop across the depletion region is given by [11, 12]:

$$\Delta V_{PV} = V_P \left(1 - \frac{J_V}{J_P} \right) \quad (3.7)$$

Assuming that at the peak bias voltage a significant portion of the core is depleted, high-speed electric field switching can be achieved, leading to high-frequency modulation of the waveguide optical transmission loss. In this situation, the RTD—OW operates as a resonant tunnelling diode electro-absorption modulator (RTD—EAM), that is, the modulation depth of the device depends mainly on the overlap between the electric field in the collector depleted region and the optical mode. In essence the increase of the magnitude of electric field is determined by the characteristics of the NDC region.

The band-edge shift described previously results in an increase of the lightwave propagation loss, that is, the peak-to-valley switching leads to a change in the intensity of the waveguide output. The modulator transmission coefficient is dependent on the applied voltage:

$$T(V) \propto \exp[-\gamma_f \cdot \alpha(V) \cdot L] \quad (3.8)$$

where $\alpha(V)$ is the waveguide core material absorption coefficient under a given bias voltage, V , L is the length of the waveguide section under electric contact, and γ_f is

the optical filling factor which corresponds to the fraction of the optical power guided in the depleted region of the waveguide.

For simplicity of analysis, neglecting the reflections and couplings losses at the facets of the waveguide, the key figure-of-merit of the modulator is modulation depth, or the modulator on/off ratio (or contrast ratio), $R_{\text{on/off}}$, due to the peak-to-valley switching, $R_{\text{on/off}} = R_{\text{VP}}$, which is given by:

$$R_{\text{on/off}} = \frac{P_{\text{out,on}}(V_{\text{on}} = V_{\text{P}})}{P_{\text{out,off}}(V_{\text{off}} = V_{\text{V}})} = \frac{T(V_{\text{P}})}{T(V_{\text{V}})} \quad (3.9)$$

In decibels,

$$R_{\text{on/off}}(\text{dB}) = 10 \log \frac{T(V_{\text{P}})}{T(V_{\text{V}})} = 4.343 \gamma_f [\alpha(V_{\text{V}}) - \alpha(V_{\text{P}})] \cdot L \quad (3.10)$$

that is:

$$R_{\text{VP}} = R_{\text{on/off}}(\text{dB}) \cong 4.343 \gamma_f \Delta \alpha \left(\frac{hc}{\lambda}, \Delta F_{\text{PV}} \right) \times L \quad (3.11)$$

where γ_f is the optical filling factor, which corresponds to the fraction of the optical power guided in the depleted region of RTD waveguide, and L is of the RTD—OW.

The magnitude of the modulation depth (or extinction ratio or the on/off ratio) can be made as large as possible by increasing the active length, L , of the RTD—OW. However, the insertion loss of the transmission (on) state also increases with L :

$$\Upsilon_{\text{in}} = \frac{P_{\text{in}} - P_{\text{on}}(V_{\text{P}})}{P_{\text{in}}} = 1 - T(V_{\text{P}}) = 1 - \exp[-\gamma_f \alpha(V_{\text{P}}) \cdot L] \quad (3.12)$$

An optimized design maximizes the extinction ratio and minimizes the insertion loss.

Another fundamental figure-of-merit in the design of electro-absorption modulators is the absorption coefficient change per unit of applied voltage:

$$\frac{\Delta \alpha}{\Delta V} = \frac{\alpha(V_{\text{off}}) - \alpha(V_{\text{on}})}{V_{\text{off}} - V_{\text{on}}} \quad (3.13)$$

These two figures-of-merit can be combined as the modulator on/off ratio (in dB) per unit of applied voltage, given by:

$$\frac{R_{\text{on/off}}(\text{dB})}{\Delta V} = 4.343 \frac{\alpha(V_{\text{V}}) - \alpha(V_{\text{P}})}{\Delta V_{\text{VP}}} \quad (3.14)$$

In our previous work, the RTD—EAM concept was demonstrated at 900 nm using the AlGaAs/GaAs material system, and at 1550 nm with InGaAlAs/InP. The quaternary compound InGaAlAs and InGaAsP lattice matched to InP can be used to implement RTD—EAM operating at wavelengths within the 1300 nm and 1550 nm optical windows, wavelengths where the standard single-mode optical fibres have zero dispersion and lowest loss, respectively. In what follows, we describe

in detail the principle of operation of an RTD—OW electro-absorption modulator and its function at optical communication windows around 900 nm and 1550 nm.

3.5 GaAs RTD—EAM operation at 900 nm

3.5.1 Device implementation and low frequency characterization

For operation spectral region around 900 nm, the RTD—OW electro-absorption modulator (RTD—EAM) structure consists of an AlGaAs/GaAs waveguide structure on semi-insulating GaAs wafers: the waveguide core is made of GaAs (Si doped $2 \times 10^{16} \text{ cm}^{-3}$); two $\text{Al}_{0.33}\text{Ga}_{0.67}\text{As}$ (Si doped $2 \times 10^{18} \text{ cm}^{-3}$) layers are employed to form the cladding layers of the waveguide. The AlAs/GaAs/AlAs DBQW was embedded in the middle of the GaAs core layer made 1 μm thick to allow easily coupling of the light in- and output (butt-coupling). The DBQW consists of a 7 nm thick GaAs quantum well sandwiched between 1.4 nm thick AlAs barriers. The detailed description and fabrication of AlGaAs/GaAs structures can be found in [10], and a diagram of a typical ridge waveguide is show in figure 3.7; also represented are the conduction band-edge E_C and refractive index RI profiles in the growth direction.

The cladding layers are made of $\text{Al}_{0.33}\text{Ga}_{0.67}\text{As}$, a direct bandgap compound alloy. The refractive index difference between the core and cladding layers is 0.224 at 900 nm, which is sufficient to obtain highly-efficient light confinement using relatively thin cladding layers. The thickness of the top cladding layer is 300 nm, twice the reciprocal of the first mode exponential decaying factor, in order to achieve a low series resistance. Because the waveguide core and the substrate have similar real refractive indices, in order to significantly reduce radiation leakage into the substrate the bottom cladding layer is twice as larger as the top cladding, i.e. 600 nm thick, therefore separating the core from the substrate and acting as an isolation layer. Figure 3.8 shows the top view of a RTD—OW die, (a), and of a packaged device, (b).

A typical current–voltage (I – V) characteristic of the AlGaAs/GaAs devices is shown in figure 3.9.

The optoelectronic characterization uses light from a *tunable* Ti:sapphire laser and includes measurements of (i) the waveguide propagation loss, determined by sequential cleaving method, (ii) the waveguide absorption band-edge, and (iii) the

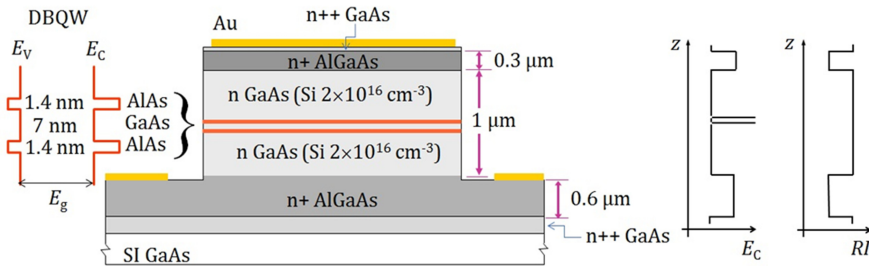


Figure 3.7. Diagram of a ridge GaAs/AlGaAs DBQW waveguide epi-layer structure; also shown are the conduction band-edge and refractive index profiles in the growth direction.

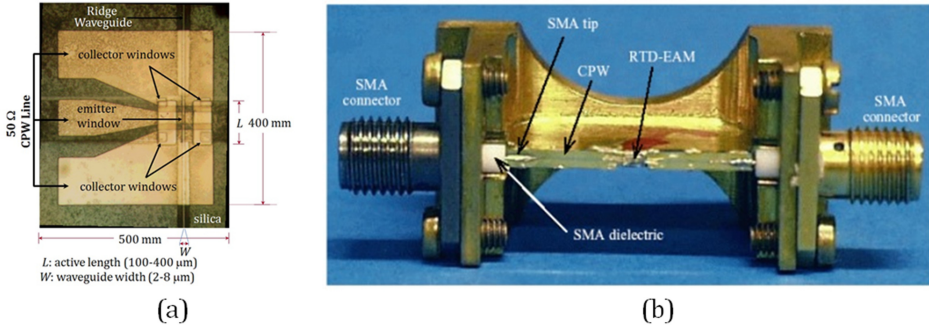


Figure 3.8. Photograph of: (a) a RTD—EAM die, here L represents the device electrical active length, which with the waveguide width defines devices active area; and (b) a connectorized prototype package device.

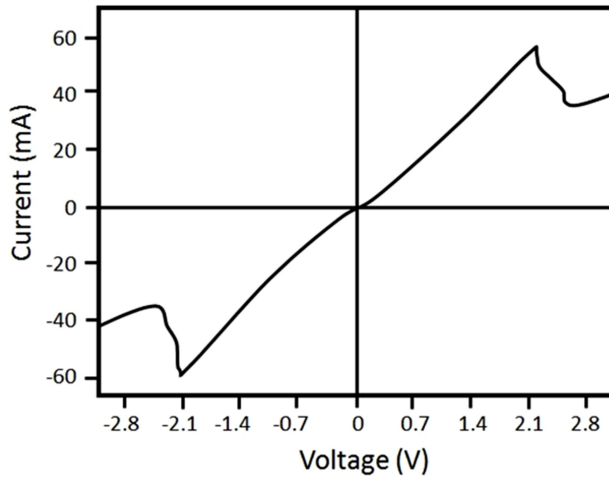


Figure 3.9. Typical full I - V characteristic AlGaAs/GaAs with a 7 nm GaAs and 1.7 nm AlAs barriers. The peak current was around 13.5 kA cm^{-2} and the peak-to-valley current ratio (PVCR) in the range 1.4–1.8. The peak voltage values were in the range 1.5 V–3 V, and the valley voltage was in the range 1.8 V–3.5 V, depending mostly on the area of the devices.

transmission spectrum. For the case of waveguides with widths above 3 μm the propagation loss at 908 nm was below 5.5 cm^{-1} . The core band-edge was around 890 nm. The RTD—EAM waveguide transmission spectra for devices with active areas around 800 μm^2 ($4 \text{ μm} \times 200 \text{ μm}$) are shown in figure 3.10 for zero bias, when biased slightly below the peak, and at the valley point. The devices were DC biased outside the NDC region in order to avoid entering in self-oscillation region. As can be seen from figure 3.10, by changing the quiescent point from the peak to the valley region, a sharp drop in the waveguide transmission at wavelengths is observed in the range 890 nm–920 nm. The respective measured Franz–Keldysh absorption band-edge wavelength shift was $\sim 12 \text{ nm}$, which compares to the 9 nm value estimated from equation (3.6), which takes in consideration the approximations used in [9].

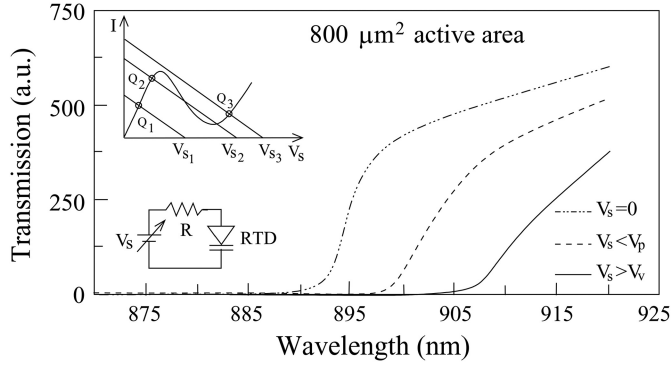


Figure 3.10. Optical transmission spectra at different DC bias points: no bias, biased slightly below the peak and in the beginning of the valley region for $4\ \mu\text{m} \times 200\ \mu\text{m}$ devices, respectively.

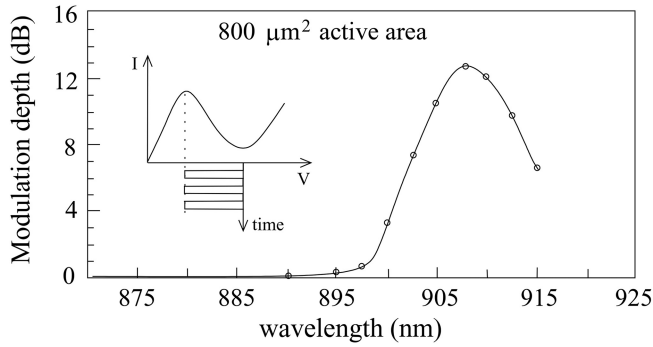
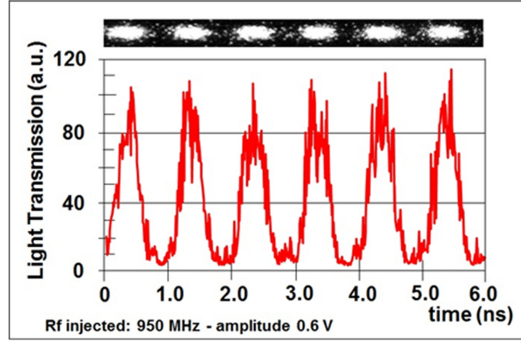


Figure 3.11. Modulation depth as function of the operating wavelength due to peak-to-valley switching induced by a square voltage waveform. The devices were DC biased in the valley region.

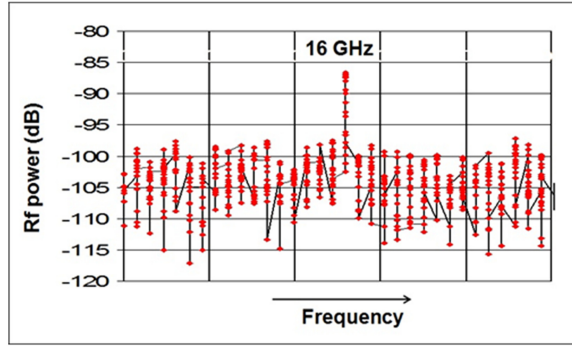
3.5.2 Optical modulation under applied external signals

In this section, we describe the RTD—OW operation as an electro-absorption modulator. The modulation depth was evaluated by driving the devices with a negative square voltage signal with a peak-to-peak amplitude slightly higher than the peak-to-valley voltage difference, ΔV_{VP} . The device DC biased in the second positive differential conductance region (where the detrimental thermal effects are less pronounced). Then an RF signal was applied through a wide bandwidth bias-T. Figure 3.11 presents the optical modulation depth as a function of the wavelength due to the transition between the two positive PDC regions. Modulation depths up to 13 dB at 908 nm were achieved. Modulation depths up to 18 dB were observed in the case of large area $4\ \mu\text{m} \times 400\ \mu\text{m}$ ridges waveguides.

Figure 3.12(a) shows examples of direct modulation at 950 MHz and 16 GHz. In both cases the amplitude of the driving signal's amplitude was $\Delta V_{VP} \cong 0.4\ \text{V}$. An optical modulation depth as large as 11 dB was achieved at 16 GHz [8, 9]. The respective 16 GHz response in the spectral domain is shown in figure 3.12(b). This provides an estimated modulation depth for the devices analysed here showing their



(a)



(b)

Figure 3.12. Direct modulation: (a) at 950 MHz; (b) at 16 GHz. The measured modulation depth was 11 dB ($\lambda = 908$ nm) at 16 GHz.

potential for high-bandwidth applications. The modulation efficiency characterized by the bandwidth-to-drive-voltage ratio, defined as the ratio of the operation bandwidth to the operating voltage ($B/\Delta V$) for a 10 dB modulation depth, was higher than 40 GHz V^{-1} [9].

3.5.3 Optical modulation enabled by self-sustained oscillations

Whenever an RTD (RTD—EAM) is made part of a resonant tank circuit, auto-oscillations are produced when DC bias point is set in the device NDC region. In this subsection, we analyse the case of RTD—EAM devices connected to a transmission line via a bias-T. This configuration forms an RTD—EAM transmission line relaxation oscillator. The inductance which is associated with the length of the transmission line together with the device capacitance defines a resonant tank circuit. When the RTD—EAM is DC biased in the NDC region self-sustained oscillations can be achieved. The frequency of oscillation can be tuned by the optical power coupled to the RTD—EAM. This behaviour enables the implementation of RTD-based optically controlled oscillators (OCOs). This frequency tuning effect is related to the creation of charge carriers in the depletion region. This reduces the

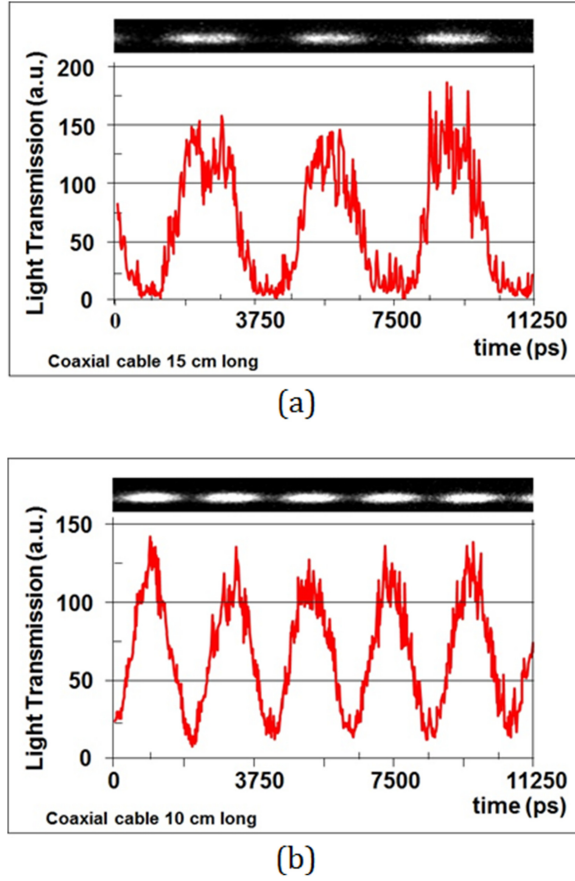


Figure 3.13. Optical responses of RTD—EAM transmission line relaxation oscillators with lines (a) 15 and (b) 10 cm long.

series resistance of the RTD—EAM and changing the operating point in the NDC region, which in turns also changes the respective impedance. Additionally, the relaxation oscillation frequency can also be tuned by changing the DC bias voltage. This changes RTD intrinsic impedance which strongly depends on the applied voltage forming a voltage-controlled oscillator (VCO).

As discussed above, when DC biased in the NDC region and connected to a bias-T through a transmission line (coaxial cable for example), the RTD—EAM self-oscillates and a modulated optical output is produced by relaxation oscillations of RTD—EAM—as shown in figure 3.13. The frequencies of oscillation are determined by the electrical length of the transmission line and the capacitance of the device.

Streak camera studies have shown that AlGaAs/GaAs RTD—EAMs under external excitation can produce pulses of light as short as 30 ps and optical modulation up 14 GHz induced by the device self-oscillation [13]. Analysis of these results suggests that the frequency was limited by the electronic packaging of the devices employed in the characterization.

The AlGaAs RTD—EAM results discussed above have potential for local area networks (LANs), primarily as devices for electrically controlling guided-wave optical signals in the wavelength range of 880 nm to 1100 nm. This includes the development of new types of waveguide intensity modulators, directional couplers and optical switches.

The high-speed self-sustained relaxation oscillations operating mode of RTD—EAMs can be applied in clock extraction circuits, for optical pulse generation and de-multiplexing in optical time division multiplexed systems.

Further, because light changes the series resistance of the resonant tank circuit then the RTD—EAM can also act as a radio frequency oscillator that has its frequency optically controlled—an optoelectronic/voltage-controlled oscillator (OVCO).

3.6 InGaAlAs/InP RTD—EAM operation at 1550 nm

3.6.1 Device structure and low frequency characterization

The AlGaAs/GaAs RTD—EAM device and its performance guided our work towards the demonstration of a conceptually similar device operating at around 1550 nm, where single-mode optical fibres have the lowest loss. Several semiconductor alloys can be used for EAMs operating at 1550 nm, such as, e.g. InGaAlAs and InGaAsP. These materials have favourable properties for high-performance RTD-based devices, such as a larger electron mobility and lower electron effective mass as compared to GaAs-based semiconductors. These properties lead to RTD—EAMs with higher performance, namely increase speed and modulation depth. The InGaAlAs/InP RTD—EAMs show larger peak current densities and peak-to-valley current ratios, and smaller peak voltage as compared to AlGaAs/GaAs RTD—EAMs. Furthermore, the quaternary system InGaAlAs is lattice matched to InP enabling operation also at 1300 nm (another telecommunications optical fibre window) [10].

Figure 3.14 shows the epi-layer structure of the InGaAlAs/InP RTD—EAM for operation at 1550 nm. It includes a waveguide structure on a highly doped InP substrate. Figure 3.14 also shows the Γ -valley conduction band profile and the refractive index profile of the waveguide. The waveguide core consists $\text{In}_{0.53}\text{Ga}_{0.42}\text{Al}_{0.05}\text{As}$ layers (refractive index of 3.56), each with a thickness of

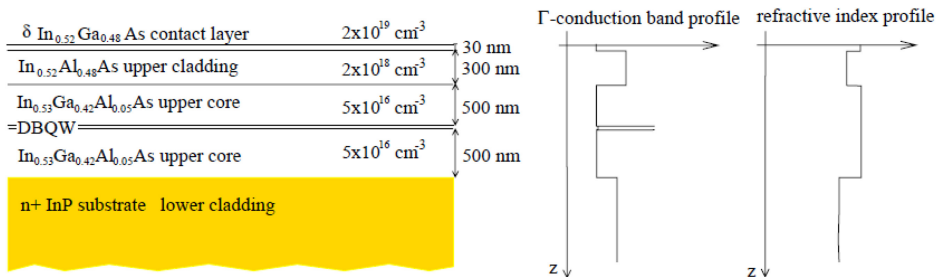


Figure 3.14. InGaAlAs/InP RTD—EAM structure, and the Γ -valley and refractive index profiles.

500 nm. The bandgap energy is 0.826 eV, corresponding to an absorption band-edge wavelength of 1500 nm for operation at 1550 nm when biased in the peak voltage. The top cladding consists of a thick 300 nm layer of $\text{In}_{0.52}\text{Al}_{0.48}\text{As}$ (refractive index of 3.24). In this configuration, the InP substrate as the bottom cladding region since the refractive index of InP at 1550 nm (3.17) is considerably lower than the refractive index of $\text{In}_{0.53}\text{Ga}_{0.42}\text{Al}_{0.05}\text{As}$. As discussed previously, the thickness of the top cladding layer thickness is 300 nm, which corresponds to approximately twice the decay length of the fundamental optical mode in this region. A delta doped InGaAs layer was included on top for ohmic contacts with lower resistance. The lower contact consisted of the highly doped InP substrate. A more detailed description of the wafer structure is given in [10].

The typical current–voltage characteristic of $4\text{ }\mu\text{m} \times 100\text{ }\mu\text{m}$ InGaAlAs/InP RTD—EAM is presented in figure 3.15, showing PVCR around 4. Most of the RTD—EAMs characterized were ridges waveguides $4\text{ }\mu\text{m}$ wide with $200\text{ }\mu\text{m}$ active lengths; the waveguides were $500\text{ }\mu\text{m}$ long in total. These devices showed valley-to-peak voltage differences ΔV_{VP} up to 0.8 V, with peak-to-valley current density differences ΔJ_{PV} higher than 10 kA cm^{-2} . (For comparison, typical GaAs/AlAs devices show differences $\Delta V_{\text{VP}} \approx 0.4\text{ V}$ and $\Delta J_{\text{PV}} \approx 5\text{ kA cm}^{-2}$.)

Impedance measurements on wafers in the frequency range 45 MHz to 18 GHz for all values of bias voltage indicates the InGaAlAs RTD—EAM small signal equivalent circuit consists of a capacitance C in parallel with a nonlinear resistor R_d , in series with a resistance R_s ; the series inductance was found to be negligible [13, 14]. The devices switching time were estimated using:

$$t_R = 4.4C_V \left(\frac{\Delta V_{\text{VP}}}{\Delta J_{\text{PV}}} \right) \quad (3.15)$$

where C_V is the device capacitance per unit area ($C_V \approx \epsilon/W_{\text{dep}}$). The average capacitance C_V and shunt resistance R_d of these devices when biased in the NDC region were estimated to be 1.5 pF and $-15\text{ }\Omega$, respectively, assuming $R_s = 3\text{ }\Omega$. This corresponds to a switching time of $t_R = 6.6\text{ ps}$ [14].

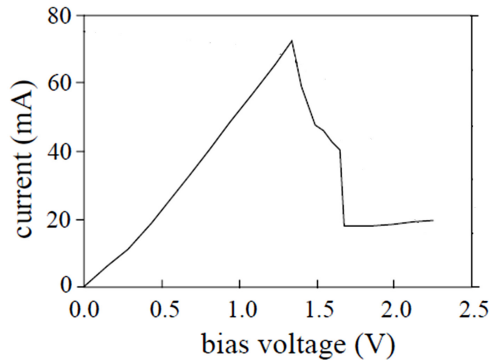


Figure 3.15. The typical current–voltage characteristic of $4\text{ }\mu\text{m} \times 100\text{ }\mu\text{m}$ InGaAlAs/InP RTD—EAM.

As previously discussed, low frequency electro-absorption response of the waveguide was characterized with only DC voltage applied and with no applied signal voltage. The DC bias voltage was either slightly below the peak voltage or in the valley region. So, the devices were biased outside the NDC region to avoid self-oscillations. Light from a Tunicas diode laser, tunable in the wavelength range 1480 nm to 1580 nm was fibre butt-coupled to the waveguide, and the light output was collected in a fibre butt-coupled arrangement. An optical power meter or a high-bandwidth photodetector was used to detect the light output. For high-speed characterization a fibre pigtailed 45 GHz bandwidth Schottky InGaAs photodetector (NewFocus) was used.

The waveguide transmission spectrum of the InGaAlAs/InP RTD—EAM was measured at various bias points, see figure 3.16. The measured band-edge shift was 43 nm, which compares well with the estimated value of 46 nm, equation (3.6). This indicates an electrical field of about 38 kV cm^{-1} is needed to achieve the observed band-edge broadening effect.

The absorption coefficient spectrum was calculated at the zero bias, when biased close to the peak and at the valley point. The results are shown in figure 3.17.

The variation of the absorption coefficient as a result of the peak-to-valley transition, $\Delta\alpha_{VP}$, was calculated and is presented in figure 3.18.

The level of electro-absorption can be estimated by calculating the change of the absorption coefficient. This is given by the peak-to-valley switching $\Delta\alpha_{VP}$ over the absorption coefficient on the on-state (in the case of the RTD—EAM, at peak bias voltage) α_P , $\Delta\alpha_{VP}/\alpha_P$ as a function of the wavelength—see figure 3.19.

The modulation depth is a function of the light wavelength as a result of the transition between the two PDC regions when a square-wave signal is applied is shown in figure 3.20. The peak-to-peak voltage was $\Delta V_{VP} \approx 0.8 \text{ V}$. The devices were DC biased in the valley region because of the smaller thermal effects in this region and to avoid self-oscillations. In this configuration, a modulation depth of 28 dB was measured for the case of a device with an active area of $800 \mu\text{m}^2$ (corresponding to ridge waveguides of $4 \mu\text{m}$ width). Impressively, these values are 10 dB larger than

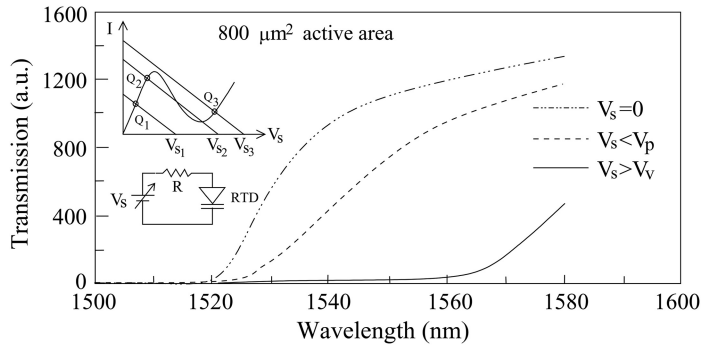


Figure 3.16. InGaAlAs RTD—EAM transmission spectra in the wavelength range 1500 nm to 1580 nm, with the applied voltage as a parameter.

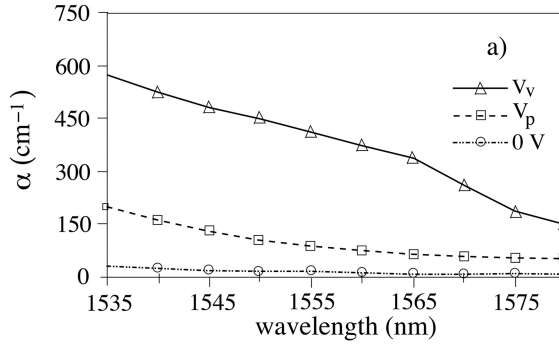


Figure 3.17. Absorption coefficient as function of the wavelength with the applied voltage as a parameter: at zero bias, and at peak and valley points.

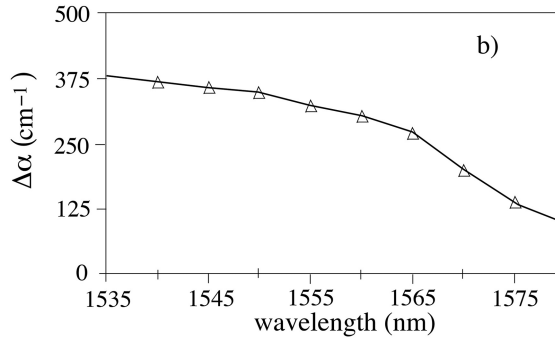


Figure 3.18. InGaAlAs RTD—EAM change in absorption coefficient spectrum in the wavelength range 1500 nm to 1580 nm when the RTD switch the voltage.

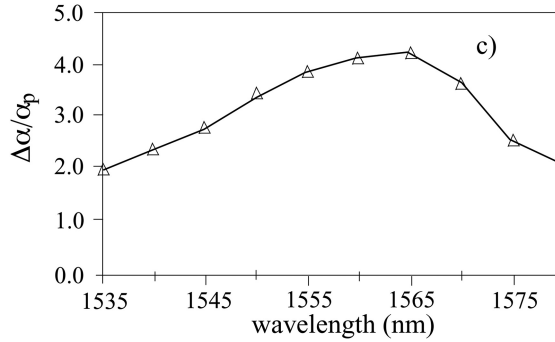


Figure 3.19. Ratio of change of absorption coefficient induced by electro-absorption, $\Delta\alpha_{vp}$, over the absorption coefficient on the on-state (at peak bias voltage) α_p .

the modulation depths measured in AlGaAs/GaAs devices. The transmissive state propagation loss was estimated to be 5 dB.

The waveguide transmission coefficient was also measured through the full I – V for some devices. Figure 3.21 presents the transmission coefficient as function of the

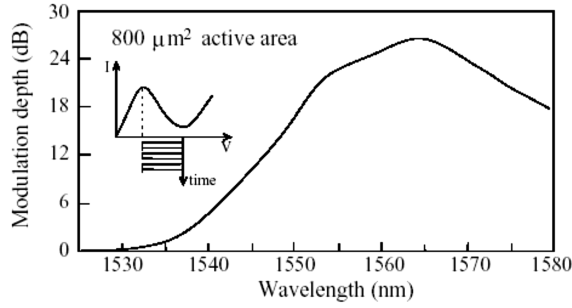


Figure 3.20. Modulation depth as function of the wavelength.

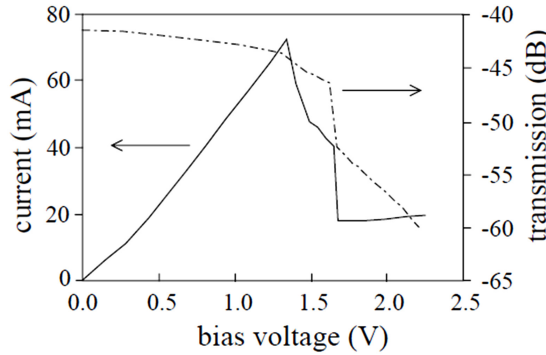


Figure 3.21. InGaAlAs/InP RTD—EAM structure, Γ -valley and refractive index profiles. Also shown, the waveguide transmission as a function of the bias voltage, where a very sharp drop in the transmission in response to a very small voltage change is seen.

applied bias voltage. The result shows a reduction of 5 dB in the transmission reduction in the case of 1 mV increase of the DC voltage at a bias point ~ 1.6 V. Effectively, this corresponds to an exceptionally large transmission change per unit of voltage.

3.6.2 RTD—EAM high-frequency characterization

First, we analyse the modulator response up to 3 GHz. Figure 3.22 plots the modulator response as function of the DC bias voltage when an RF voltage signal at 3 GHz is applied (the amplitude of the RF signal is used as a tunable parameter). The I - V characteristic of the RTD—EAM is also shown in figure 3.22. The modulator high-frequency response increased ~ 15 dB when the device DC bias changed from the peak to the valley region, even for RF driving signals at amplitudes as low as 50 mV. This is a clear indication that the modulator can achieve impressive modulation depths even in the cases of extremely low amplitude voltage signals.

In order to test the RTD—EAM at frequencies $\gg 10$ GHz, a microwave synthesized signal generator with a maximum output of +20 dBm and a maximum frequency 26 GHz was employed. The modulator response in the case of a 26 GHz

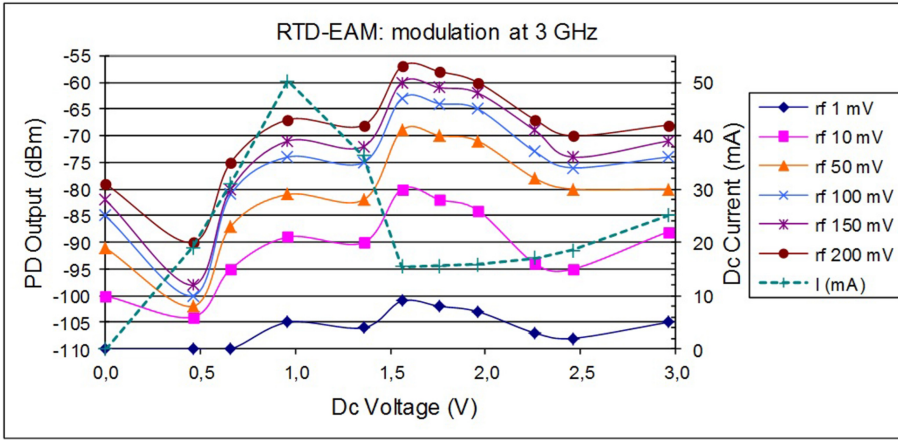


Figure 3.22. Modulator response as a function of the DC bias voltage when driven by 3 GHz rf signals, with injected amplitude as a parameter.

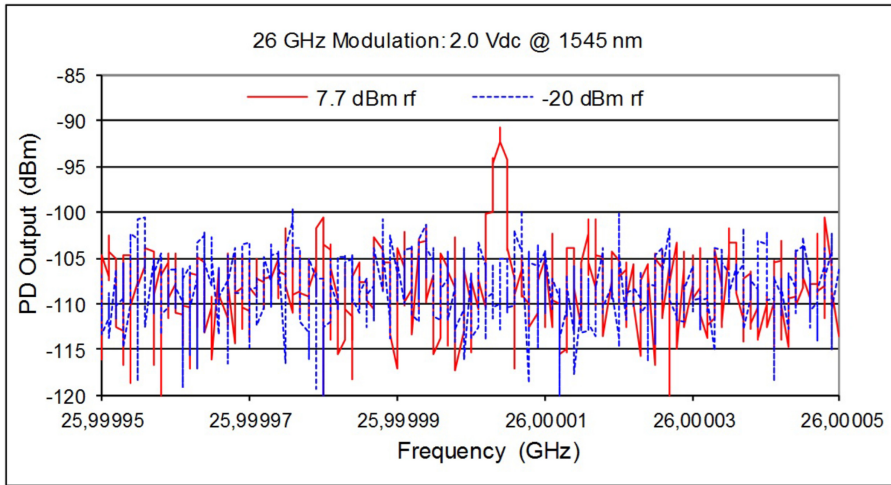


Figure 3.23. Spectra of the 26 GHz photo-detected signals for modulator driving power of -20 dBm and +7.7 dBm, respectively.

driving signal at 7.7 dBm power levels is shown in figure 3.23. The photo-detected power increases more than 10 dB when the driving rf power increases from -20 dBm to +7.7 dBm. This is a clear indication that the device is capable of reaching modulation extinction ratios larger than 10 dB for low-power driving signals, typically lower than 10 mW, as consequence result of the built-in electrical amplification effect of RTD—EAM devices. The amplification effect (due to intrinsic NDC of RTD—EAMs) can reduce substantially the rf power required to achieve large modulation depths, e.g. 10 dB. This on-chip amplification can eliminate the need of external rf amplifiers usually required to drive EAMs [1],

and therefore providing practical solutions for future high-speed and low-power modulators.

3.7 Conclusions

In this chapter, we covered the RTD—EAM and how the nonlinear current–voltage characteristic of the RTD can be used to rapidly and efficiently switch the electric field in an optical waveguide. The electric field changes the optical absorption of the waveguide via the Franz–Keldysh effect (see figure 2.8 and equation (2.11)) so that photon energies just below the bandgap energy become strongly absorbed. The waveguide EAM has been used in optical communication applications to encode digital information on to an optical signal—the advantage of incorporating the RTD is that this function can be implemented with much less rf power applied to the device.

The RTD—EAM concept was first realised in AlGaAs material system and was then transferred to an InP material system that had its bandgap tuned to the optical communications ‘c’ band, wavelengths around 1550 nm. This InP RTD—EAM device demonstrated capability at frequencies up to 26 GHz with low drive radio frequency drive powers of around 10 mW—when DC biased to the NDC region the incorporating the RTD with the EAM waveguide device obviated the need for an external rf amplifier and considerable reduced system complexity.

The intrinsic high speed of the tunnelling process in RTDs, which has already allow the demonstration of fundamental oscillations over 1.9 THz [15], and the possibility of combination with other optoelectronic and electronic devices makes the RTD—EAM a device with high potential for the next generation of information technology systems, in particular for more energy efficient high-speed communication and signal processing implementations.

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Resonant Tunneling Diode Photonics

Devices and applications

Charlie Ironside, Bruno Romeira and José Figueiredo

Chapter 4

Resonant tunnelling diode—photodetector (RTD—PD)

4.1 Introduction

A photodetector is an optoelectronic device that absorbs light at its input followed by signal conversion to an electrical output. They are widely used in a variety of applications ranging from imaging to optical communications, particularly semiconductor-based photodetectors can be described as follows: if photons have an energy larger than the bandgap energy of the semiconductor, light absorption creates electron–hole pairs that increase the semiconductor electrical conductivity and thereby increases the current flowing in an external circuit that provides a voltage signal [1].

In typical semiconductor photodetectors the light sensitive region is a *pn* junction or a low doped or intrinsic semiconductor layer embedded within a *pn* junction (*pin* junction). The built-in electric field of a *pn* junction separates the photogenerated electron and hole charges followed by their extraction to an external circuit. There are mainly two modes of operation of semiconductor photodetectors: photoconductive and photovoltaic. In the case of the photoconductive mode discussed in this chapter, an external voltage is applied to create a high electric field in the light sensitive region, which makes the photogenerated charges travel away from this region to the contact regions at high speed. In this case no voltage or only a small voltage is applied and the photogenerated carriers drive an external circuit producing an external voltage. The magnitude of the photogenerated current or voltage is proportional to the number of incident photons. In addition to the basic photodetection mechanism, some photodetectors include an electrical amplification section [2].

The PIN photodiode (PIN—PD) and avalanche photodiodes (APDs) are well-established as semiconductor photodetectors. APDs are advanced PIN—PD

structures that include extra semiconductor layers to provide internal gain as the result of a carrier (electron) multiplication process with a relatively noisy mechanism [3].

A particularly important application for photodetectors is optical communication systems. A high-speed light-wave communication system includes (i) a transmitter optoelectronic circuit employing a near-infrared laser diode (and for very high data bit rates (>10 GBs) an optical modulator), where the light-wave carrier is modulated by an electric waveform embedding the encoded information signal to be transmitted; (ii) a transmission media, such as an optical fibre, eventually utilising in-line optical amplification, and (iii) a receiver circuit based on a high-speed photodetector that recovers the encoded information signal from the light-wave to the electric domain.

Photodetectors receive the transmitted optical pulses and convert them, in practical situations using as low energy loss as possible, to electric pulses that can be then used by other electronic devices connected at the receiving end node. In general, the electrical output of a photodetector requires an amplification stage. The performance of the information signal recovering process is highly dependent on the noise level at the receiver. Therefore, the main requirements for these detectors include large responsivity (generally defined as the ratio of the photocurrent and the optical power incident on the photodetector), high sensitivity (defined as minimum amount of optical power incident on the photodetector that produces a distinguished signal, a signal to noise ratio equal to 1), low noise, wide-bandwidth, high reliability, and low cost [4, 5].

In state-of-the-art high-speed optical communications, typically PIN—PD or APD detectors are used. Both the PIN—PDs and APDs are reversed bias devices (that is, the cathode is made positive with respect to the anode) and operate well within the photoconductive mode.

In PIN—PDs the dark current is very small (typically below <1 nA), and the total current is directly proportional to the light irradiance. In these cases, to maximise the sensitivity of the device (that is, to minimise the amount of optical power being detectable) the dark current must be as low as possible. When APDs are employed to provide internal gain it is crucial to minimise the noise.

Considering that double barrier quantum well (DBQW) resonant tunnelling diodes (RTDs) are structures fully compatible with III—V semiconductor optoelectronic devices, it is expected that novel RTD-based detectors can be implemented taking advantage of the DBQW nanostructure. In this chapter, a new photodetector structure based on the integration of a DBQW within or close to the light absorbing region of unipolar photoconductive detectors is introduced. These novel devices, known as resonant tunnelling diode photodetectors (RTD—PDs), are capable of realising both photodetection and amplification functionalities. They are similar to the DBQW—RTD devices discussed in the previous chapter. However, for the photodetection application they include at least one light absorbing layer larger than the DBQW (typically with a thickness between 200 nm—1000 nm). It is expected that the RTD—PD can be designed to take advantage of RTD high-speed, wide-bandwidth electrical gain, low noise, while providing an effective optical

responsivity larger than currently available photoconductive detectors. Due to its built-in wide-bandwidth electrical gain they are being considered as potential candidates for high-speed optical-to-electrical transducers for future light-wave and radio-over-fibre communication systems, and as single-photon detectors for quantum-based technologies. There has been world-wide research effort in exploring the potential of RTD—PDs [6–10] and more generally the potential of photo-detectors that utilise negative differential conductance (NDC).

Compared with PIN—PDs and APDs, RTD—PDs could achieve much higher bandwidth-efficiency product due to the high-speed and considerably larger effective responsivity due to their high nonlinear photocurrent-irradiance characteristic. Moreover, the RTD—PDs electrical amplification mechanism is, in principle, a much lower noisy process when compared with APDs carrier (electron) multiplication mechanism, since it is a consequence of the resonant tunnelling current dependence on the applied voltage which leads to an NDC region. Overall, as potential advantages over the APD and PIN—PD, the RTD—PD offers a much lower operating voltage (typically around or below 1 V), higher bandwidth and lower noise levels.

Noteworthy, an advantage of the RTD—PD compared to APD is that the electrical gain noise is not subject to random impact ionisation processes but to general thermal considerations and the bandwidth is limited usually by parasitic components as is the case with the RTD oscillators which have been demonstrated to operate up to 1.92 THz [11].

The addition of light sensitive layers to the RTD conventional epi-layers enables the realization of radio-frequency oscillators capable of operating in the terahertz band that can be controlled by optical signals and this has obvious applications in signal processing and telecommunications. Due to their large wide-bandwidth NDC, RTD—PDs do offer other modes of operation when combined hybrid or monolithically with other electronic and optoelectronic devices. Among others there are the voltage controlled optoelectronic oscillators and optical controlled microwave oscillators (see chapter 6), with applications in telecommunication data links as optical-to-wireless converters and as local oscillators.

In this chapter, we will discuss the implementation and principle of operation of RTD—PD designed to operate as a high-speed optical-to-electrical transducer for optical communication systems [12]. The RTD—PD is a recent and promising invention that has yet to establish itself that provides at some time photodetection and electrical amplification.

4.2 RTD—PD concept and principle of operation

An RTD—PD can be understood as a modification of a unipolar semiconductor photoconductive detector, see figure 4.1(a). The RTD—PD corresponds to the monolithic integration of a DBQW with a unipolar (isotype) $N-n-n/i-n-N$ photoconductive epi-layer structure, as schematically shown in figure 4.1(b). A photoconductive detector is a device where the absorption of light increases the number of free charge carriers (electron and holes) which raises its overall electrical

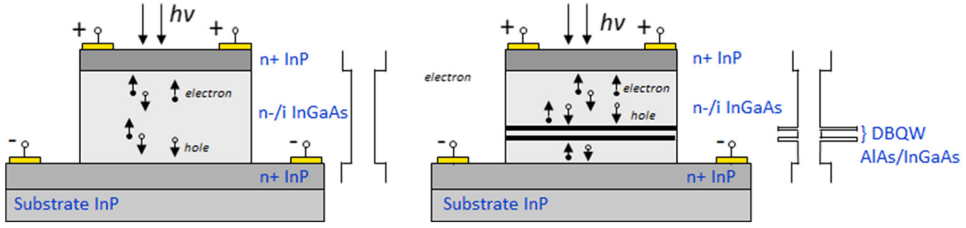


Figure 4.1. (a) Schematic of physical layout of InGaAs/InP unipolar semiconductor photoconductive detector, (b) Schematic representation of a photoconductive detector incorporating a DBQW structure consisting of AlAs barriers and InGaAs quantum well.

conductivity. The total current flowing in the external circuit is the sum of the dark current (current that is generated in the absence of light) and the additional photocurrent. To a first order, for a given light spectral distribution the photocurrent is linearly proportional to the light irradiance (power received by the optical window surface per unit area).

The RTD—PD operation principle is as follows: as the bias voltage increases there is a progressive band bending, more pronounced in the collector side of the structure, that lowers the collector barrier relative to the emitter. This results in an increase of the number of electrons in the emitter side being able to cross the structure through the DBQW first quantised energy level (through the charge transport mechanism known as resonant tunnelling), this is the first positive differential conductance (PDC) region of device I – V characteristic. In the first PDC region the current through device follows the bias voltage until the voltage reaches a certain value that leads to a current local maximum; this happens when the DBQW first quantised level aligns with the emitter Fermi-level. Increasing the voltage above this value leads to a sharp decrease of the current till it reaches a local minimum at the valley voltage, because of the reduction of the carrier transmission across the DBQW. This current decrease over a given voltage range corresponds to the device I – V NDC region. Above the valley current other mechanisms make the current increase again—the beginning of the second PDC region.

When a light-wave with energy above the light absorbing layers bandgap energy is incident in the structure it modulates the charge transport mechanisms, increasing the devices overall conductivity and modulating the internal built-in electrical field. Its impact will depend on the location of the light absorbing layer that gives the highest contribution to the photo-response. Let us first consider the case of RTD—PD structures where the light absorbing layer is in the collector side, figure 4.2(a). Under illumination we observe an overall shift of the DC I – V characteristic to lower voltage (also termed as voltage red shift) as shown in figure 4.2(b). The impact of the light absorption can be described as follows: due to the electric field induced by the applied voltage, the photogenerated electron–hole pairs become separated by the built-in electric field induced by the applied voltage, with the photogenerated electrons drifting towards the collector contact while photogenerated holes move towards the emitter, figure 4.2(c). However, these photogenerated holes tend to accumulate in the collector region adjacent to the DBQW, leading to an electric field

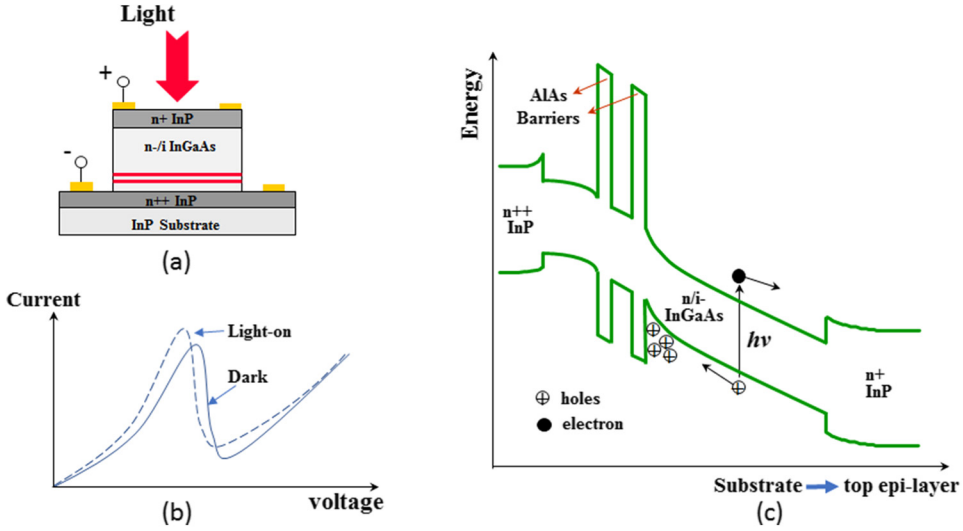


Figure 4.2. (a) Physical layout of a RTD—PD with the light absorbing region on the collector side. (b) Representation of the DC I - V characteristic light-induced voltage shift to lower voltage. (c) Schematic representation of the RTD—PD conduction and valence bands profile under bias, and of the principle of operation of the RTD—PD, showing the effects of light absorption: accumulation of holes in the collector region adjacent to the DBQW and electrons drifting towards the collector contact.

enhancement across the DBQW and the collector depleted region adjacent to the DBQW, making the resonance be achieved at a lower voltage. This leads to the overall shift of the DC I - V curve under illumination to lower voltage (also termed as voltage red shift), as shown in figure 4.2 which gives a schematic representation of the principles of operation of an RTD—PD for the case where the light absorption layer is located between the DBQW and the collector contact. The additional photocurrent which adds to the DC (dark) current of the device is the result of the device conductivity. The voltage global shift and the enhancement of the built-in electrical field across the DBQW and the adjacent depleted collector region are mainly attributed to the accumulation of the photogenerated holes in the region adjacent to the collector barrier, as represented schematically in figure 4.2(c).

The photocurrent (difference of the current under illumination and the dark current) at a given bias voltage is proportional to the incident optical power. The overall photocurrent is a nonlinear function of the applied voltage since the device photoinduced conductivity enhancement is a nonlinear function of the applied bias. The amount of voltage shift depends on the applied voltage and on the optical irradiance being absorbed by the device. Moreover, due to the built-in electrical field dependence on the charge distribution the device responsivity (the ratio between the photocurrent and the optical power absorbed) becomes a function of the bias voltage and the light irradiance.

In generic RTD—PD structures there is at least one low doped or intrinsic semiconductor layer sensitive to light adjacent to the DBQW. Figure 4.3 shows schematic diagrams of: (a) the conventional RTD, where the thickness of the spacer

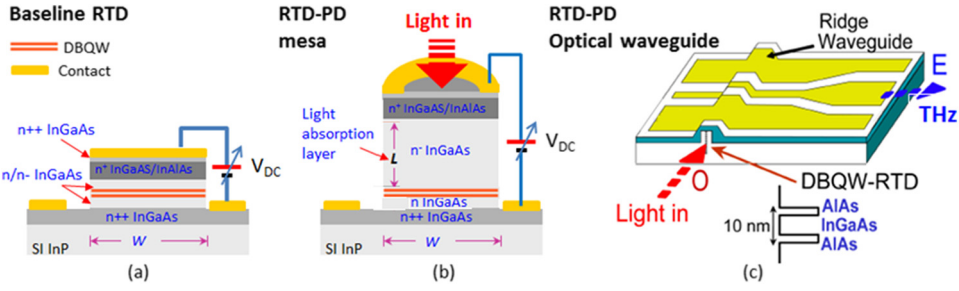


Figure 4.3. Schematic diagrams of (a) conventional RTD structure, (b) RTD—PD mesa, and (c) optical waveguide implementations.

layers between the DBQW and the high-doped emitter and collector contact regions are just few tens of nanometres (it may include additional undoped spacer layers adjacent to the DBQW that helps to suppress diffusion of charge carriers from the doped layers to the AlAs barriers, preventing dopants contaminating the barriers); (b) the RTD—PD in the mesa configuration, where the DBQW and the intentional added light absorbing region corresponding to the wide low doped/intrinsic layer several are embedded between higher bandgap (lower refractive index) materials; and (c) RTD—PD optical waveguide which employs an epi-layer structure similar to the mesa's RTD—PDs, where the higher bandgap (lower refractive index) layers are the cladding layers of the waveguide.

In the mesa configuration, figure 4.3(b), an optical window is created on the top contact layer to allow the light to reach the photosensitive regions of the device. In the RTD—PD waveguide configuration, figure 4.3(c), the light is butt-coupled to the waveguide with the help of a lens or/and an optical fibre and propagates along the photo-absorbing (higher refractive index) layers in the plane of the DBQW.

The RTD—PD operation takes advantage of electrical field magnitude high-nonlinearities in the direction of the current flow induced by the carrier filtering effect of the DBQW in the N - n -DBQW- n - N structure. It also provides an electrical amplification mechanism by adding to the current–voltage characteristic of the N - n - i - n - N photoconductive structure an NDC region that can persist at room temperature from DC up to few terahertz. These effects give rise to a significant enhancement of the photodetector responsivity that is both dependent on the irradiance level and applied voltage.

In summary, the RTD—PD can be operated in three modes:

- (i) 'passive' mode (non-oscillating—DC biased in the first PDC region away from the NDC), which operates as a classical PD but taking advantage of the RTD—PD nonlinear optoelectronic response that leads to an effective responsivity much larger than the p - i - n photodiodes;
- (ii) excitable mode, where a pulsed light signal can shift the RTD—PD quiescent point to the NDC region, making it to oscillate during the time the light is on;

- (iii) in the oscillation mode when DC biased in the NDC region, with the light signal modulating the RTD—PD oscillation frequency and at some extent its RF power.

When biased close to or within the NDC, intensity modulated light can be used to induce the quiescent operation point to move to the NDC, putting the RTD—PD in the oscillation mode, or to induce peak-to-valley switching which gives rise to a current change that (ideally) can be equivalent to the peak-to-valley current difference that could potentially lead to absolute responsivities values of tens of A W^{-1} at irradiance levels of the order of few hundred of microwatts to few milliwatts.

4.3 RTD—PD implementation and characterisation

The RTD—PD characterisation employs the main photodetection standard metrics (it is worth pointing out that these photodetection standard metrics are defined for linear systems—see for example [13]), that include responsivity (which is defined as the ratio of generated photocurrent and the incident (or sometimes absorbed) optical power, determined in the linear response region [4, 5, 14]), represented by $\mathcal{R}$; sensitivity (the smallest photo-signal that can be detected related to the noise or noise equivalent power); and the photodetection bandwidth. In this section we present the performance of InGaAs/InP RTD—PD devices in terms of responsivity and discuss how it compares to PIN—PDs and APDs.

As will become clear from the discussion the nonlinear nature of the RTD—PD and APD these photodetection standard metrics for linear system metrics must be used with care when designing systems that employ RTD—PDs and APDs. The objective is to design RTD—PD devices whose performance overcomes PIN—PDs and APDs metrics required for high-speed optical communication systems.

Optical telecommunications photodetectors are optimised mainly with respect to high bandwidth-quantum efficiency product (or gain bandwidth product), sensitivity (lowest detectable light level) and high-output saturation power (responsivity). The efficiency of a photodetector as an optical-electrical converter is normally characterised by its responsivity. In an ideal photodetector every photon produces one electron-hole pair which is transferred instantly and without losses to the contact electrodes of the photodiode, giving rise to the light-generated current, that is, the quantum efficiency of the ideal photodetector is equal to one.

The responsivity of a photodetector is usually defined for the steady state. The term *responsivity* should not be confused with *sensitivity*; the latter is the lowest detectable light level, which is typically determined by the noise level of the receiver and is significantly influenced by the required detection bandwidth. From the point of view of particularly the RTD—PD the concept of responsivity has a limited range of applicability since it is only valid to the linear part of output current versus input optical power region of the response curve and the RTD—PD, as we shall see, has important nonlinear regions in this curve.

As mentioned, the responsivity $\mathcal{R}$ was originally defined with linear systems in mind. Because the RTD—PD I – V characteristic is highly nonlinear making it a

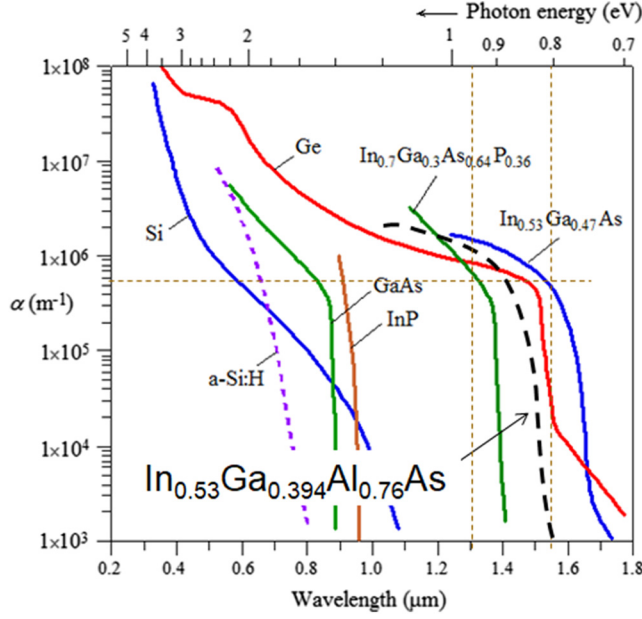


Figure 4.4. Absorption coefficient α as a function of the wavelength λ for several semiconductor materials [4, 5]; spectral response of a $\text{In}_{0.53}\text{Ga}_{0.47}\text{As}$ photodetector [15] is highlighted.

nonlinear system, it is expected that the linear definition of $\mathcal{R}$ does not fit precisely with the operation of an RTD—PD. So, in general, the RTD—PD does not have a single value for $\mathcal{R}(\lambda)$ that covers the full voltage range of its operation for a given wavelength λ . It is expected that RTD—PD responsivity will depend on the bias voltage V and the wavelength λ and the power of the optical signal, that is $\mathcal{R} = \mathcal{R}(\lambda, V, P_{\text{ph}})$.

The RTD—PD responsivity is calculated in the steady state (or stationary mode), which corresponds to the photodetector response to an incident continuous light-wave (constant intensity optical signal). The RTD—PD steady (stationary) state responsivity $\mathcal{R}(\lambda, V)$, measured in A W^{-1} , is calculated as [14]:

$$\mathcal{R}_G(\lambda, V) = \left| \frac{I_{\text{ph}}(\lambda, V) - I_{\text{dark}}(V)}{P_{\text{ph}}(\lambda)} \right| = G(V, P_{\text{ph}}) \eta(\lambda) \frac{e\lambda}{hc} \quad (4.1)$$

where $I_{\text{ph}}(\lambda, V) - I_{\text{dark}}(V)$ is the current difference under illumination and dark conditions, and $P_{\text{ph}}(\lambda)$ is the continuous-wave optical power incident on the device; $G(V, P_{\text{ph}})$ accounts for the internal current gain effect (due to electron–hole pair generation, built-in electrical gain, and/or impact ionisation mechanisms, or other phenomena not yet investigated), $\eta(\lambda)$ is the quantum efficiency which is a function of the optical material properties (absorption coefficient and refractive index) and also of the thickness of the absorbing region, and e , h , c are the elementary charge, Planck’s constant and the speed of light in vacuum, respectively. Figure 4.4 shows the absorption coefficient spectra $\alpha(\lambda)$ for the majority of the semiconductor

materials employed in high-speed photodetectors, including the InGaAs ternary and InGaAlAs quaternary semiconductor compounds lattice matched to InP.

For a PIN—PD the responsivity, $\mathcal{R}(\lambda, V)$, in the linear part of its light-current curve is given by [14]:

$$\mathcal{R}(\lambda, V) = \frac{I_{\text{ph}}(\lambda) - I_{\text{dark}}}{P_{\text{ph}}(\lambda)} \approx \frac{I_{\text{ph}}(\lambda)}{P_{\text{ph}}(\lambda)} = \eta(\lambda) \frac{e\lambda}{hc} = \eta(\lambda) \frac{\lambda \text{ (}\mu\text{m)}}{1.24} \quad (4.2)$$

where $I_{\text{ph}}(\lambda)$ is the photoinduced current ($I_{\text{ph}}(\lambda) \gg I_{\text{dark}}$). So, for example, at a wavelength of 1.24 μm the maximum value possible for the responsivity $\mathcal{R}$ is 1AW, which requires the maximum quantum efficiency, $\eta = 1$.

The effective responsivity of a PIN—PD can be increased by incorporating electrical gain. However, the electrical gain must be incorporated with minimum noise and maximum bandwidth for optimal photodetector performance. For APDs the electrical gain noise (or so-called excess noise) [16] arises from the random nature of impact ionisation and the bandwidth of APD electrical gain is determined by the avalanched build-up time [14] and the so-called dead time [17] over which the avalanche build-up takes place and both are subject to random processes. For photodetector devices with electrical gain, G , such as APDs, the effective responsivity is as follows:

$$\mathcal{R}_G(\lambda) = G(V, P_{\text{ph}}) \eta(\lambda) \frac{e\lambda}{hc} = G(V, P_{\text{ph}}) \eta(\lambda) \frac{\lambda \text{ (}\mu\text{m)}}{1.24} \quad (4.3)$$

$G(V, P_{\text{ph}})$ is the average number of electrons produced in the external circuit per photogenerated electron–hole pair.

In the case of RTD—PDs, the position of the light absorbing layer(s) relatively to the DBQW will determine the effect of the light-generated electron–hole pairs on its I – V characteristic and on its optoelectronic response. Several effects and loss mechanisms lead to reduction of responsivity. The photodetection quantum efficiency value is reduced by the limited absorption in finite absorption layers thickness, residual optical reflection of the incident light at semiconductor surface, losses due to fibre-chip coupling, losses in internal contact layers and carrier recombination (before the contacts are reached). Moreover, the photocurrent generated in the high-frequency range operation will be smaller than the stationary state value because of bandwidth limiting effects such as RC time and transit time constraints.

We have considered several InGaAlAs/InP RTD—PD epi-layer structures aiming to demonstrate a new class of photodetectors that takes advantage of the non-linearity and, in particular, the NDC of the DBQW—RTD structures.

Table 4.1 describes in detail the epi-layer structure of InGaAs/AlAs RTD—PD devices comprising a 500 nm low doped InGaAs layer for light absorption within the wavelength range 1100 nm to 1600 nm.

Figure 4.5 shows the physical layout of InGaAs/InP RTD—PD and the epi-layer conduction and valence band profiles. Mesa RTD—PD devices 10 $\mu\text{m} \times 10 \mu\text{m}$, with

Table 4.1. Epi-layer structure of an InGaAs/InP RTD—PD designed to operate in the wavelength range of 1100 nm–1600 nm.

Layer #	Thickness (nm)	Material	Doping (cm ⁻³)	Description
1	45	In _{0.53} Ga _{0.47} As	3 × 10 ¹⁹ ; Si	Collector
2	80	In _{0.53} Ga _{0.47} As	3 × 10 ¹⁸ ; Si	Collector
3	500	In _{0.53} Ga _{0.47} As	5 × 10 ¹⁶ ; Si	Absorption
4	2	In _{0.53} Ga _{0.47} As	Undoped	Spacer
5	1.7	AlAs	Undoped	Barrier
6	4.7	In _{0.53} Ga _{0.47} As	Undoped	Well
7	1.7	AlAs	Undoped	Barrier
8	2	In _{0.53} Ga _{0.47} As	Undoped	Spacer
9	20	In _{0.53} Ga _{0.47} As	2 × 10 ¹⁶ ; Si	Spacer
10	80	In _{0.53} Ga _{0.47} As	3 × 10 ¹⁹ ; Si	Emitter
11	400	In _{0.53} Ga _{0.47} As	3 × 10 ¹⁹ ; Si	Emitter
12	200	InP	Undoped	Buffer
		InP Substrate		

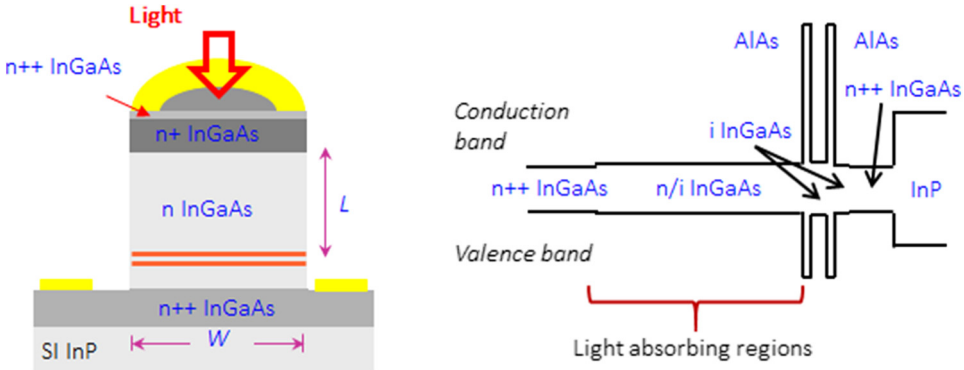


Figure 4.5. Physical layout of InGaAs/InP RTD—PD mesas with a light absorbing layer 500 nm thick, capable of light detection within the wavelength range of 110 nm to 1600 nm. Also shown is the representation of the device conduction and valence band profiles at zero bias.

5 $\mu\text{m} \times 5 \mu\text{m}$ optical window, were implemented. A single mode optical fibre was used to couple light through an optical window on the top of the mesa.

The implemented devices showed peak current densities around $0.3 \text{ mA } \mu\text{m}^{-2}$ and peak-to-valley current ratios (PVCr) close to 1.5. Under dark conditions, the device's NDC region was between 1.1 V and 1.4 V. The device's stationary photodetection responses were characterised using unmodulated light-waves with wavelengths 1310 nm and 1550 nm for three optical power levels: dark, 0.1 mW, and 1 mW. Figure 4.6(a) show the typical the I - V characteristics under dark conditions and under illumination at 1310 nm.

It is clearly noticeable in figure 4.6(a) that under illumination the I - V characteristic shifts to lower voltage (a voltage red shift), with the light-induced voltage red shift being proportional to the injected optical power. It is also evident that the device photo-response is nonlinear, with the responsivity being a function of the bias voltage and the optical power. The measured responsivity is about 10 A W^{-1} at bias close to the peak voltage and for injected light power of 0.1 mW. The responsivity dropped to 5 A W^{-1} when the optical power was increased to 1 mW. The results confirm that the responsivity is a highly nonlinear function of the optical power.

Figure 4.6(b) shows the device response to 1550 nm wavelength light measured using the same conditions and power levels. The maximum measured responsivity was 3 A W^{-1} for 0.1 mW and 2 A W^{-1} for 1 mW optical powers, respectively. The lower values of responsivity obtained at 1550 nm are mainly associated to the lower absorption coefficient of the InGaAs at 1550 nm, $0.68 \mu\text{m}^{-1}$ at 1550 nm, when compared with $1.16 \mu\text{m}^{-1}$ at 1310 nm.

We have also investigated the photodetection response of InGaAlAs/InP RTD—PD epi-layers that can also be operated as electro-absorption modulators at around 1550 nm. The corresponding epi-layer is described in table 4.2

The InGaAlAs/InP RTD—PD epi-layer includes two 500 nm InGaAlAs (7% of Al) quaternary compound lattice matched to InP each side of the DBQW and two

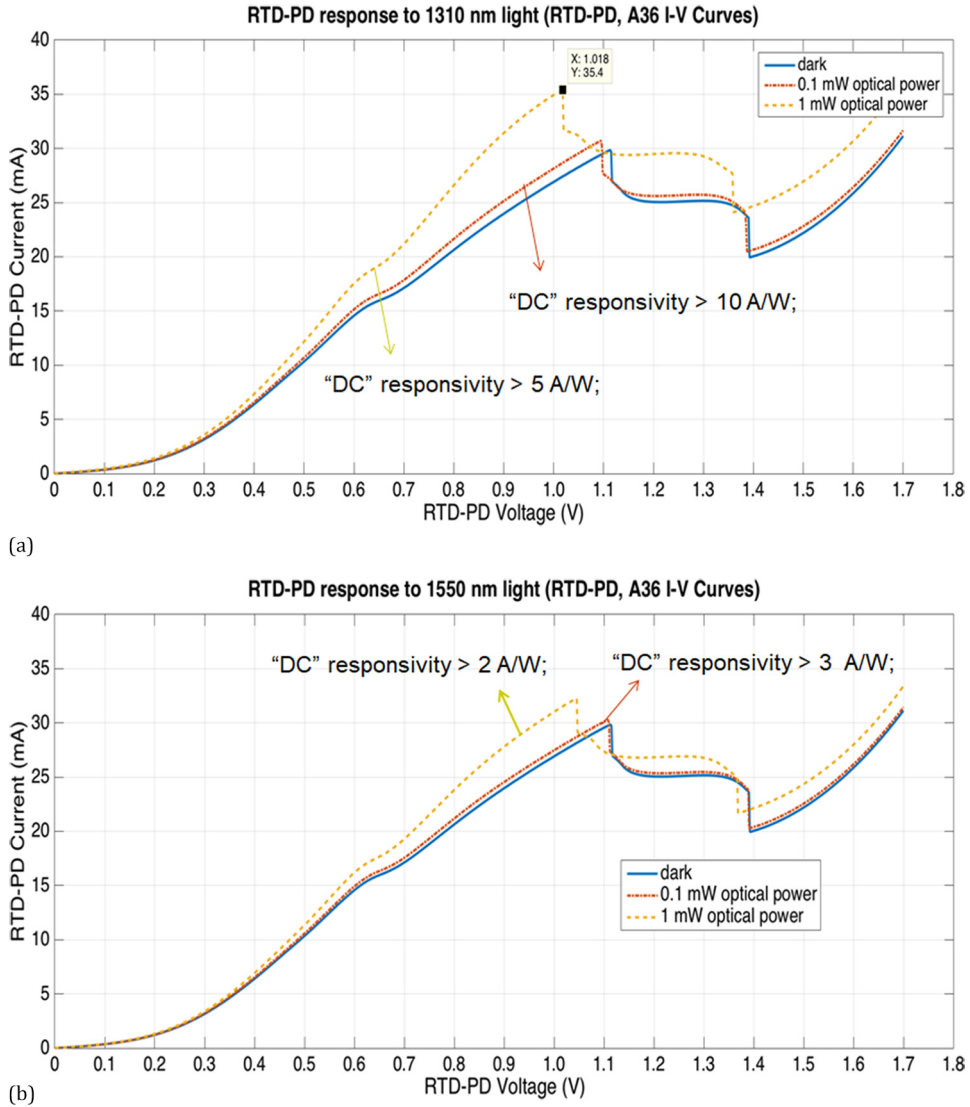


Figure 4.6. Measured current–voltage response of mesa InGaAs/InP RTD–PD illuminated at wavelengths (a) 1310 nm, and (b) at 1550 nm.

extra InAlAs cladding layers in the case of the RTD–PD waveguide implementation. Mesa RTD–PD devices $25\ \mu\text{m} \times 25\ \mu\text{m}$ in size, with $15\ \mu\text{m} \times 15\ \mu\text{m}$ optical windows, were fabricated.

Figure 4.7 shows a schematic of the physical layout of InGaAs/InP RTD–PD; also shown is the representation of the conduction and valence band profiles of the device at zero bias.

Figure 4.8(a) presents the typical InGaAlAs/InP RTD–PD forward bias I – V characteristic at dark conditions (blue line), and when illuminated with 1310 nm

Table 4.2. Epi-layer structure of InGaAlAs/InP RTD—PDs designed to operate in the wavelength range of 1100 nm–1560 nm.

Layer number	Thickness (nm)	Material	Doping(cm^{-3})	Description
1	100	$\text{In}_{0.532}\text{Ga}_{0.468}\text{In}$	$2 \times 10^{19} \text{ Si}$	Collector
2	300	$\text{In}_{0.521}\text{Ga}_{0.479}\text{In}$	$2 \times 10^{18} \text{ Si}$	Collector
3	500	$\text{In}_{0.53}\text{Ga}_{0.394}\text{Al}_{0.076}\text{In}$	$5 \times 10^{16} \text{ Si}$	Absorption
4	2	$\text{In}_{0.53}\text{Ga}_{0.394}\text{Al}_{0.076}\text{In}$	Undoped	Spacer
5	1.7	AlAs	Undoped	Barrier
6	5.7	$\text{In}_{0.532}\text{Ga}_{0.468}\text{In}$	Undoped	Well
7	1.7	AlAs	Undoped	Barrier
8	2	$\text{In}_{0.53}\text{Ga}_{0.394}\text{Al}_{0.076}\text{In}$	Undoped	Spacer
9	500	$\text{In}_{0.53}\text{Ga}_{0.394}\text{Al}_{0.076}\text{In}$	$5 \times 10^{16} \text{ Si}$	Absorption
10	300	$\text{In}_{0.521}\text{Ga}_{0.479}\text{In}$	$2 \times 10^{18} \text{ Si}$	Emitter
11	500	$\text{In}_{0.532}\text{Ga}_{0.468}\text{In}$	$2 \times 10^{19} \text{ Si}$	Emitter
12	100	$\text{In}_{0.521}\text{Ga}_{0.479}\text{In}$	Undoped	Spacer
13	20	InP	Undoped	Buffer
		InP Substrate		

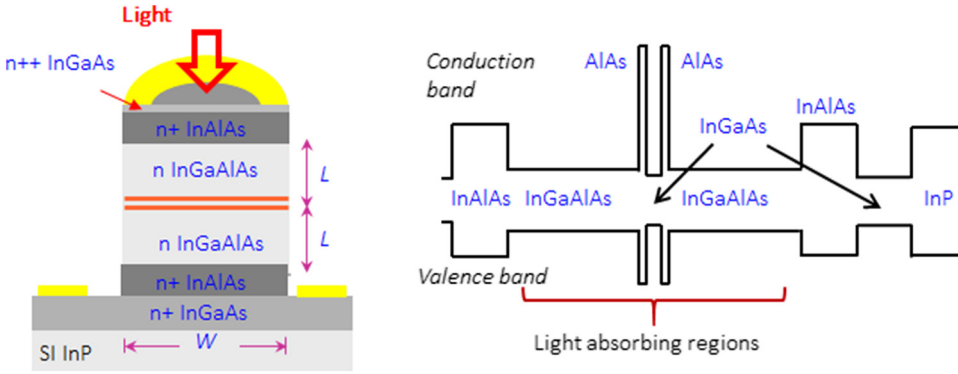


Figure 4.7. Physical layout of InGaAs/InP RTD—PD mesas with light absorbing layers 500 nm thick both sides of the DBQW, capable to light detection within the wavelength range of 110 nm to 1560 nm. Also shown is the representation of the bottom of the conduction energy and top of the valence band energy profiles versus distance for the device at zero bias.

light for optical powers: 0.25 mW (red line), 0.5 mW (yellow line) and 1 mW (purple line). The measured responsivities were 28.8 A W^{-1} and 15.1 A W^{-1} on the first and second PDC, respectively, for 0.25 mW illumination; 24.4 A W^{-1} and 13.5 A W^{-1} on the first and second PDC, respectively, for 0.5 mW; 19.7 A W^{-1} and 10.8 A W^{-1} on the first and second PDC, respectively, for 1 mW. The light-induced voltage blue shift was quantified as the light voltage sensitivity measured as the peak voltage shift per unit of optical power, measured in V W^{-1} . The measured light voltage sensitivity was 204.8 V W^{-1} and 408.4 V W^{-1} on the first and second PDC, respectively, for 0.25 mW; 144.4 V W^{-1} and 318 V W^{-1} on the first and second PDC, respectively, for 0.5 mW; 102.2 V W^{-1} and 262 V W^{-1} on the first and second PDC, respectively, for 1 mW.

Figure 4.8(b) shows the InGaAlAs/InP RTD—PD photo-response under reverse bias: I – V characteristics under dark conditions (blue line), and under illumination for 0.25 mW (red line), 0.5 mW (yellow line) and 1 mW (purple line). The measured responsivities were 10.8 A W^{-1} and 2.4 A W^{-1} on the first and second PDC, respectively, at 0.25 mW; 10 A W^{-1} and 3.6 A W^{-1} on the first and second PDC, respectively, at 0.5 mW; 8.5 A W^{-1} and 4.2 A W^{-1} on the first and second PDC, respectively, for 1 mW illumination. The lower responsivity values can be attributed to the lower effectiveness of the applied voltage to induce the bands bending in order to achieve the resonant condition: the peak voltage occurs at higher voltage (well above 1 V, against 0.7 V for forward bias) with the peak current 50% higher when compared with forward bias value.

The light-induced voltage blue shift was quantified as follows: 204 V W^{-1} and 180 V W^{-1} on the first and second PDC, respectively, at 0.25 mW, 138 V W^{-1} and 216 V W^{-1} on the first and second PDC, respectively, for 0.5 mW, 84 V W^{-1} on the first PDC when illuminated with 1 mW.

As shown in figure 4.9 for symmetric structures the illumination induces an overall I – V curve shift towards higher voltage (voltage blue shift) that increases with

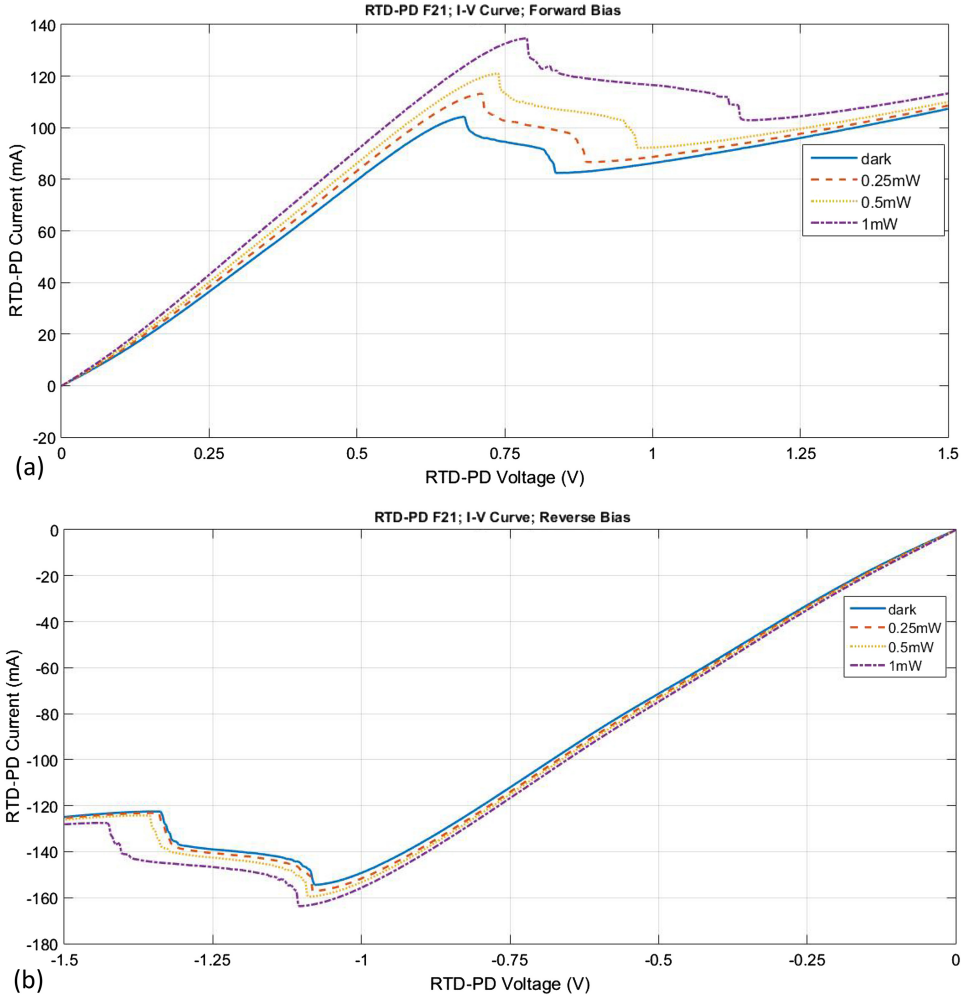


Figure 4.8. Measured current–voltage response of mesa InGaAlAs/InP RTD—PD illuminated at a wavelength of 1310 nm, under (a) forward and (b) reverse bias respectively.

optical power. This I – V characteristic global blue shift is mainly attributed to the stronger impact of the photogenerated carriers in the emitter, when compared with the effect of the accumulation of holes in the collector region adjacent to the DBQW. Additional studies will be required in order to quantify the impact of the locations and the size of the light absorbing layers relatively to the DBQW.

The photodetection response of an InGaAs/InP RTD—PD to a 1310 nm modulated light at 1 GHz is shown in figure 4.9 with the optical power as a parameter. The inset shows the scope photo-detected signal. The highest responsivity obtained for these devices was 3.3 A W^{-1} . If we consider a PIN—PD with an equivalent intrinsic layer 1 μm thick with a quantum efficiency equal to 1 (not a

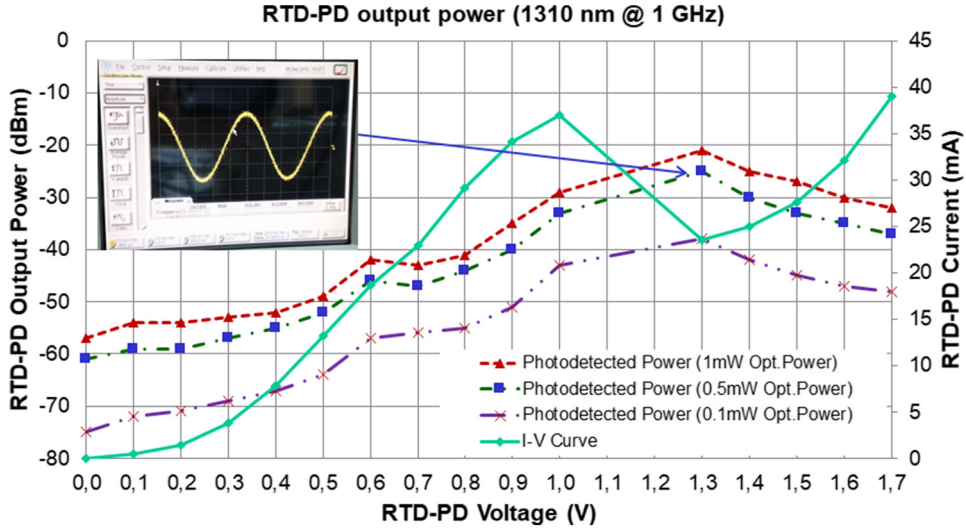


Figure 4.9. InGaAs/InP RTD—PD response to a 1310 nm light-wave modulated at 1 GHz with the optical power as a parameter. Inset shows the scope photo-detected signal.

realistic value, but quantum efficiencies up to 0.7 easily are achievable), the ideal responsivity would be below 1 A W^{-1} within the 1100 nm–1600 nm wavelength range. Here we are not considering the Fresnel losses that can be minimised and are assuming a coupling efficiency equal to 1 (i.e. all arriving light is coupled to the photodetector).

The above results demonstrate that the RTD—PD can be designed to operate as a photodetector with larger values of responsivity than the PIN—PD. It was also demonstrated [8, 18] that RTD—PD structures can reach much higher values of responsivity when operating at very low levels of irradiance (well below microwatt). With optimised structures and maximising coupling of light into the absorption region of the RTD—PD very large responsivities as high as $\mathcal{R}_G = 3.1 \times 10^4 \text{ A W}^{-1}$ have been observed with a sensitivity of several pA per photon [19].

As shown previously the RTD—PD photo-response has a highly nonlinear component which depends on the DC bias voltage and on the injected optical power so that $\mathcal{R}_G$ becomes dependent on the applied bias voltage and optical power. When operated in the first PDC well below the peak (passive mode) where the bias voltage is set to maximise the value of $\mathcal{R}_G$, the effective responsivity can be well above the APD, as mentioned in the previous paragraph. For high-speed telecommunication applications operating around a high responsivity for an APD is considered to be around $\mathcal{R}_G \sim 1 \text{ A W}^{-1}$ [20].

Another interesting feature of RTD—PD devices is that when the bias voltage is set below the peak but not far from the NDC region a high-frequency pulsed optical signal can induce fast changes in the conductivity of the device giving rise to a high-frequency pulsed current flowing in the external circuit. This high-frequency pulsed current can be used to directly drive other electronic and optoelectronic components

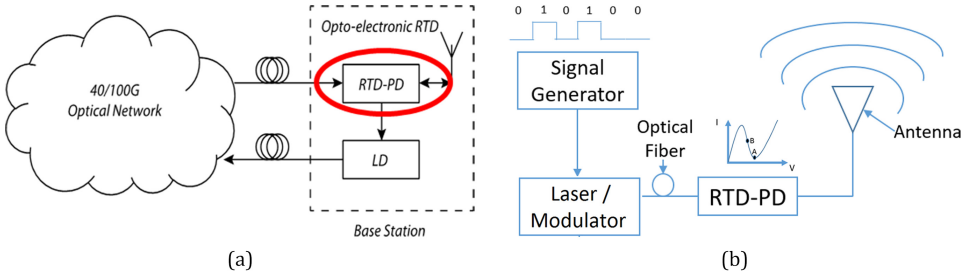


Figure 4.10. (a) Schematic representations of RTD—PD application on radio-over-fibre wireless communication systems. The RTD—PD is configured as photodetector with built-in gain and as an electronic oscillator. The phase and/or the amplitude of the RTD—PD oscillator output can be controlled by both an electrical and an optical signal. (b) Schematic of optically controlled RTD—PD configured as an optical-to-wireless data link.

such as semiconductor lasers [21]. This capability is being used in novel optoelectronic systems such as optoelectronic oscillators [22], optical/wireless transducers links [23], and neuromorphic optoelectronic circuits.

Next, we discuss the RTD—PD operation as optical-to-microwave transceiver and optical controlled oscillator.

4.4 Optical controlled high-frequency RTD—PD oscillators

High-speed photonic-microwave interfaces are a key technology for the coming generations of higher capacity radio-over-fibre based wireless communication systems, 5G and above. The requirement for tens of GBps per user data rates needs a concomitant increase in carrier frequency resulting in systems using carrier frequencies of hundreds of gigahertz (GHz). Although synthesised multiplier-based systems can be used to achieve these high frequencies, they require many stages of multiplication resulting in high power consumption and large size. An alternative approach is the use of advanced solid-state oscillators with fundamental frequency in the terahertz range as the carrier frequency. As mentioned previously, RTDs are currently the basis of the highest frequency solid-state oscillator and therefore they have the potential to satisfy the requirements of the next generations of radio-over-fibre based wireless communication systems. Indeed, RTD—PDs are thought to be potential candidates for the low-cost integration of radio and fibre networks. In this section we discuss the RTD—PD oscillator capabilities for optical-to-microwave transceivers in wireless data links [24, 25]. The concept of the RTD—PD optical-to-wireless data link is illustrated in figure 4.10. The RTD—PD is configured as an electronic oscillator. The phase and/or the amplitude of the RTD oscillator output is controlled by an optical signal.

Generically speaking, high-frequency RTD—PD circuits can operate in three modes, figure 4.11: (a) non-oscillating (steady state); (b) excitable; and (c) oscillating, depending on the operating point. When biased in the PDC regions of the I — V curve sufficiently away of the NDC, figure 4.11(a), the RTD—PD shows similar functionality as PIN—PD and APD devices, converting the intensity modulated optical signal into an electrical signal that resembles the electrical signal used to

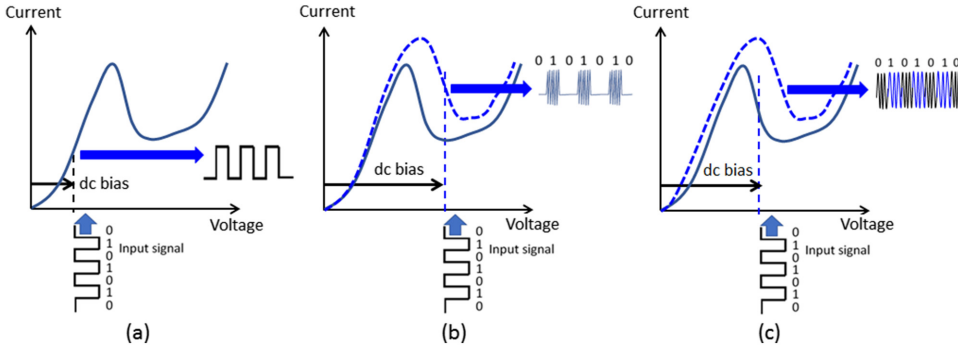


Figure 4.11. Schematic representation of the modes of operation of a RTD—PD: (a) non-oscillation mode, when biased in the PDC sections of the I — V curve; (b) excitable mode, when biased close to but outside the NDC region (here in the second PDC region); and (c) in the oscillation mode, when biased on the NDC.

modulate the light-wave signal. When the RTD—PD is biased in the PDC regions but close to the NDC figure 4.11(b), an intensity modulated optical signal incoming with an intensity above a given threshold shifts quiescent point of RTD—PD oscillator to the NDC of the I — V under illumination, resulting in burst of oscillations lasting the time of the duration the high intensity signal, as schematically shown in figure 4.11. In the case the bias point is within the NDC of the dark I — V , the RTD—PD is permanently oscillating. The presence of an intensity modulated signal will modulate the frequency of RTD—PD oscillations. In all three operation modes the optical-to-electrical conversion will benefit from the built-in gain due to the device high nonlinearity(ies).

As described in [24], an RTD oscillator circuit comprising an DBQW AlAs/InGaAs/AlAs RTD—PD with a InGaAlAs optical absorption layer was used as the RTD—PD for the optical wireless link. The bandgap of the InGaAlAs alloy was adjusted so that light at optical communications wavelengths (1300 nm and 1500 nm) was strongly absorbed in the light absorption layers. The RTD—PD circuit oscillation frequency was set to 34 GHz given by the intrinsic RTD—PD capacitance and the inductance of the CPW line. When DC biased to the appropriate voltage, in this case in the range ~ 0.7 – 1.05 V the RTD—PD oscillated at 34 GHz and could be amplitude modulated, phase modulated and frequency modulated by combination of the applied optical signal and the exact setting of the DC bias voltage.

In figure 4.12 we show the output signal in the time domain of the RTD—PD amplitude modulated at a data rate of 80 Mbit s^{-1} . It was also possible to demonstrate quadrature phase shift keying (QPSK) with a similar set-up and to demonstrate data rates up to 200 Mbit s^{-1} .

Subsequent work [25] has operated with higher carrier frequencies up to 79 GHz and data rates of 50 Mbits s^{-1} .

As described in [17] injected modulated RF signals and intensity modulated optical signals were able to shift the bias operation point of RTD—PD circuit with fundamental frequency of oscillation at 79 GHz but biased at the end of the first PDC region, just before the NDC region, from the steady state (PDC region) to the

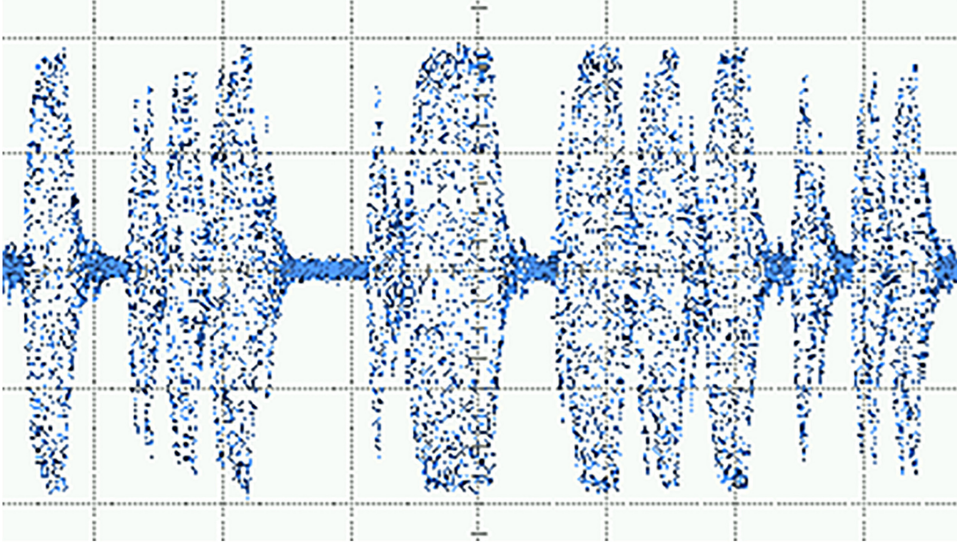


Figure 4.12. The output signal in the time domain from the RTD—PD showing data patterns at 80 Mbit s^{-1} amplitude modulating a RF carrier wave at 34 GHz.

oscillatory state (NDC region) to generate a mm-wave signal that transfers the input data signal onto an RF carrier.

Importantly, this technique can be scaled to higher data rates and higher carrier frequencies and when used with integrated antenna, can result in very low-cost, high performance flexible electronic, optical and radio interfaces.

4.5 Design guidelines of high photodetection bandwidth RTD—PDs

For high-speed digital optical communications systems and many other applications, it is required that the individual components, such as photodetectors, have a frequency response $H(\omega)$ with a 3 dB frequency (or 3 dB bandwidth) well above the corresponding maximum foreseen data bit rate for the system. The 3 dB bandwidth is the frequency at which the magnitude of the response has fallen off by 3 dB. For optical telecommunications, the bandwidth metric is crucial [26] as it largely determines the data rates which are now in the range 10 Gb s^{-1} – 100 Gb s^{-1} [27] and of course in a single channel telecommunications system the overall bandwidth is limited by the bandwidth of the smallest bandwidth component. The bandwidth of an optical receiver is limited by the physical process in the photodetector and by how the parasitic, inductance, resistance and capacitance contribute to the time constants of the external circuit.

As mentioned above, the photodetector response typically falls off for intensity modulated signal frequencies above some detection frequency. There are three factors that limit the speed of a photodetector: diffusion of carriers, drift transit time in the depletion region, and capacitance of the depletion region. The slowest of the three processes is the diffusion of carriers to the high-electric field depletion region

from outside that region. To minimise this slow effect, carriers should be generated near or in the depletion region. Reducing the ratio between the diffusion length and the drift length of the device results in a greater proportion of the generated current being carried by the faster drift process. Diffusion of the carriers generated outside the depletion region is a relatively slow process—caused by the optical absorption in the diffusion regions outside of the high-field depletion region (diffusion current can last as long as the carrier lifetime), they lead to the formation of a long tail in the impulse response of the photodiode, which results in a low-frequency falloff in the frequency response. The diffusion time τ_{diff} for carriers to diffuse a distance L can be estimated using equation below:

$$\tau_{\text{diff}} = \frac{L^2}{D} \quad (4.4)$$

where D is the minority carrier-diffusion coefficient. For InGaAs electron and hole diffusion coefficients are of the order of $150 \text{ cm}^2 \text{ s}^{-1}$ and $5.5 \text{ cm}^2 \text{ s}^{-1}$ [4], respectively, which, for $L = 500 \text{ nm}$, gives a hole diffusion time τ_{diff} around 227 ps.

The 3 dB electrical frequency (frequency where the collected current has decreased by 3 dB) of a photoconductive detector limited by the carrier-diffusion time can be estimated using the following equation [28]:

$$f_{3 \text{ dBdiff}} \approx \frac{3.5}{2\pi} \frac{1}{\tau_{\text{diff}}} \quad (4.5)$$

Using the value of $\tau_{\text{diff}} = 227 \text{ ps}$ it gives a 3 dB frequency $f_{3 \text{ dBdiff}} \sim 3.3 \text{ GHz}$. For a high-speed photodetector, this diffusion mechanism has to be eliminated; for example, by reducing the photo-generation of carriers outside the depletion layer through an improved design of the device structure, such as including advanced heterojunctions.

The second process, the transit time, is the time required for the carriers to drift across the depletion region and get swept out of the device. For higher frequency operation the photogenerated carriers, particularly the electron, transit time across the structure must be small, which implies that L must be minimised and the photogenerated carrier velocities, particularly the electron velocity, in the semiconductor must be maximised. With sufficient bias, electrons and holes will drift at their saturation velocities. The 3 dB electrical frequency of a photoconductive detector limited by the carrier-transit time can be calculated using [28]:

$$f_{3 \text{ dBtt}} \approx \frac{3.5v}{2\pi L} \quad (4.6)$$

where $v^{-4} = \frac{1}{2}v_e^{-4} + \frac{1}{2}v_h^{-4}$. The parameters v_e and v_h are the electron and hole saturation velocities, respectively. For very high bandwidth (thin) photodetectors, the bandwidth-efficiency product $f_{3 \text{ dBtt}} \times \eta_{\text{ph}}$:

$$f_{3 \text{ dBtt}} \times \eta_{\text{ph}} \approx \frac{3.5v}{2\pi L} \kappa(1 - R_{\text{ref}})\alpha(\lambda) \quad (4.7)$$

The expected intrinsic carrier-transit time limited 3 dB frequency $f_{3\text{ dBtt}}$ of the RTD—PD mesa structure shown in figure 4.7, with a light absorbing layer of $L = 500$ nm, is of the order of 60 GHz, using $v = 5.3 \times 10^6$ cm s⁻¹ (from electron and hole velocities in InGaAs, $v_e = 6.5 \times 10^6$ cm s⁻¹ and $v_h = 4.8 \times 10^6$ cm s⁻¹, respectively [28]). It corresponds to a transit time of around 10 ps.

Here we have assumed the absorbing layer thickness is equal to the depleted (high electric field) intrinsic/low-doped layer thickness. This assumption is quite ambitious and does not take into account that the depletion region is bias dependent, while the intrinsic layer thickness is fixed by the epi-layer design.

For mesa type photodetectors (that is, structures where the absorbing layer is perpendicular to the illumination), several compromises in design have to be considered with respect to the thickness of the active photon-absorbing layer. The ideal bandwidth-quantum efficiency product for perpendicular illuminated photodetectors depends on the operating wavelength. A large absorption layer thickness is needed to achieve high quantum efficiency and hence high responsivity. However, a large absorbing layer results in longer carrier drift time. As a numerical reference and assuming single-pass absorption, it is estimated that for an absorbing coefficient of the order of $1\text{ }\mu\text{m}^{-1}$ the ideal bandwidth-efficiency product will be below 20 GHz and 35 GHz in the third and second optical communication windows, respectively [28].

Lastly, the capacitance of the device determines its RC time constant (here, R represents the device/circuit load resistance, usually $50\text{ }\Omega$). To maximise a photodiode response, the transit time and diffusion time are typically designed to be comparable but smaller than the RC time constant. For instance, given the saturation velocity for InGaAs, a 10 ps transit time, for example, requires that the depletion layer be $0.5\text{ }\mu\text{m}$ thicker at most. For a comparable RC time constant in a $50\text{ }\Omega$ system, the capacitance must be less than 200 fF. Assuming the capacitance C can be estimated using the parallel plate approximation,

$$C \approx \epsilon \frac{A}{d} \quad (4.8)$$

where d is the dielectric thickness (here the depleted region), here $d = 0.5\text{ }\mu\text{m}$, A is the active area of the device, and ϵ is the dielectric relative electrical permittivity ($\epsilon = 13$ for InGaAs). Considering the numerical values above, we find the device active area must be less than $870\text{ }\mu\text{m}^2$ (equivalent to a disc with a diameter smaller than $35\text{ }\mu\text{m}$). However, it is highly unlikely that the RTD—PD 500 nm absorbing layer will become completely depleted unless it is designed as an unintentionally doped or very low-doped layer. If we assume a 100 nm collector depleted region, the maximum area must be lower than $170\text{ }\mu\text{m}^2$ (equivalent to a circular optical window diameter smaller than $15\text{ }\mu\text{m}$).

For photoconductors [14] the recombination times τ of electrons and holes dominate many aspects of the design of large bandwidth devices. Reverse biased PIN photodiodes tend to be faster than photoconductors because they have a wide intrinsic section inserted between the p and n regions allowing large built-in fields to be obtained that separate and rapidly sweep the electrons and holes towards the

electrodes. This wide intrinsic section leads to a lower capacitance, thereby a smaller RC time constant, and to the suppression of relatively slow diffusion processes compared to faster drift processes in transporting the photogenerated carriers to the electrodes.

In RTD—PD the most crucial and limiting parameter seems to be the lifetime of photogenerated and accumulated charge carriers, which sets upper bounds for the device sensitivity and operation frequency [29].

A combination of the responsivity, $\mathcal{R}_G$ and the noise in a photodetector with gain determines the sensitivity that is the smallest amount of energy that a photodetector can register with a given signal to noise ratio; if the $\text{SNR} = 1$ then this is the minimum detectable signal.

For RTD—PD the excess noise factor, F , has not been extensively studied. There has been some work on the shot noise of the current flowing through an RTD and there is evidence that as a consequence of the quantum tunnelling process the current shot noise through a RTD is suppressed [30]. The resonant tunnelling process has been employed in quantum dots for single-photon detectors albeit at cryogenic temperatures [31].

4.6 Excitability detectors

A key concept from nonlinear dynamics that provides insights into the operation of both RTD—PDs and APDs is excitability [32]. The concept of excitability originates from biological systems and the operation of neurons—and it provides useful insights for computing systems that emulate neural networks [33]. Both RTD—PDs and APDs [34, 35] are types of excitability detectors; that is detectors that can deliver a large pulse output for a small trigger input but then have a dead time while the external circuit recharges the device so that it can then deliver another output pulse when it receives the next trigger input.

We can make an analogy between bush fires, which are excitable events triggered by lightning strikes, and excitable photodetectors. The bush fire requires combustible material to build-up over a timescale of several months and then the trigger event, the lightning, ignites the fire and that is followed by dead time in which lightning cannot cause a fire because there is not enough combustible material. So the analogy is that using excitable photon detectors like RTD—PDs and APDs is similar to using a bush fire to detect lightning. We can expect false positives; not all bush fires are caused by lightning and false negatives not all lightning strikes produce fires. Not all pulses from an excitable photodetector are caused by photons and some photons do not produce output pulses—this adds to the noise from the excitable photon detector. The bandwidth is limited by the length of the pulse (how long the fire lasts) and the dead time (how long it takes for the combustible material to build up). The best takeaway from this analogy is that the simple linear models used for most photodetectors with internal amplification are incomplete and a more complete understanding of excitable photodetectors requires an analysis based on nonlinear dynamics—see chapter 6.

4.7 Conclusions

In this chapter, we covered the RTD—PD, where the highly nonlinear current–voltage (I – V) response of the DBQW—RTD is employed to amplify a weak incoming optical signal. The RTD—PD is in a similar class of detectors like the APD that can be described using concepts from nonlinear dynamics such as excitability. The nonlinear dynamics of RTD photonic devices is more comprehensively covered in chapter 6. The highly nonlinear response of the RTD—PD makes it suitable for digital communication systems but some of the metrics from linear photodetectors have to be used carefully when describing RTD—PD (and indeed APDs).

Several groups world-wide have worked on RTDs or more generally tunnelling devices as integrated amplifiers in photodetectors. Although in this chapter we concentrated on the review of RTD—PD operating in near-infrared for optical communication systems and radio-over-fibre wireless communication systems, but the concept has general applicability and has been used for mid-infrared detectors [10, 36] and large amplification factors have been reported—up to 10^5 excess electrons for one absorbed photon [10].

Because the RTD has a large bandwidth NDC, it can easily be configured as a high-frequency electronic oscillator. When light absorption layers are included the RTD—PD can be used as a high-frequency electronic oscillator controlled by an optical signal. This has been shown to be the basis of optical-to-wireless data links operating with carrier frequencies of up to 79 GHz and data rates of 200 Mbit s^{−1}. This approach clearly has the potential for high carrier frequencies and increased data rates because the RTDs have been shown to be the key active component in electronic oscillators operating at up to 810 GHz with powers of 180 μW [37] and at 1.92 THz with powers of 0.4 μW [11].

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Resonant Tunneling Diode Photonics

Devices and applications

Charlie Ironside, Bruno Romeira and José Figueiredo

Chapter 5

Resonant tunnelling diode—laser diode (RTD—LD)

5.1 Introduction

For III–V semiconductor devices, there is considerable convergence between electronics and optoelectronics in terms of epitaxial layer material growth, device fabrication techniques, and device layout. This has prompted investigations into integrating semiconductor light emitting devices with purely electronic devices to produce optoelectronic integrated circuits (OEICs) with expectations of emulating the gains in functionality, speed and compactness that have been so significant in the development of integrated silicon electronics. There has been work on RTD light emitting diode integration [1], RTD—laser diode integration [2–4], Gunn diode and laser integration [5] and transistor laser integration [6].

As noted in chapter 4, the RTD—PD has large enough electrical gain (or responsivity) that it can directly drive another device such as a semiconductor laser and this has been an important feature for the development of the RTD-based optoelectronic oscillator (OEO) the subject of chapter 7. In this chapter we discuss the development of the RTD—laser diode (RTD—LD) with particular emphasis on it as a component of the OEO.

5.2 Hybrid RTD—laser circuit

The hybrid RTD—LD consists of separate laser diode and RTD chips electrically coupled via a short (~2 mm) transmission line as illustrated in figure 5.1.

Compared to a fully integrated device, an advantage of the hybrid approach is that it is possible to separately characterise the current–voltage (I – V) properties of the laser and RTD chips and then combine them to investigate how the I – V curve changes—as illustrated in figure 5.2.

The RTD I – V curves illustrated in figure 5.2 have a region indicated as the NDR (related to the NDC) region; in this flat plateau region the hybrid circuit is in fact

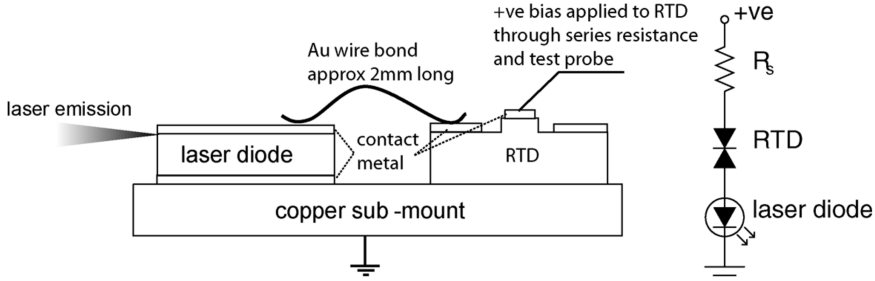


Figure 5.1. Illustration of the RTD—LD hybrid integration module. The RTD and laser dies are attached to the copper sub-mount with silver epoxy, a wire bond is made between them using a wedge bonder. On the right is a schematic lumped element circuit representation of the device [2].

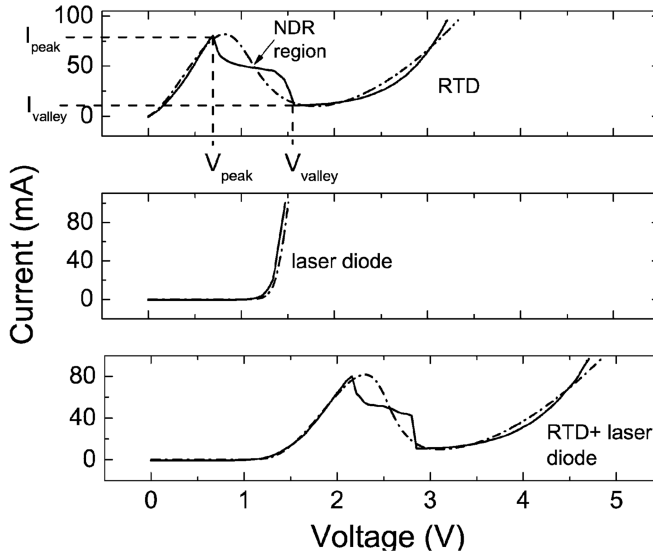


Figure 5.2. The I - V characteristics of the RTD—LD; top the RTD chip, middle the laser diode chip and bottom the hybrid RTD plus laser circuit. The solid lines are the measured data and the dashed lines are fitted from a PSPICE model as reported in [2].

oscillating and the oscillation frequency is determined by the value of the external parasitic components that are introduced by the bias circuit.

5.2.1 Model of the RTD—LD hybrid circuit

Initially the RTD—LD was modelled using an established circuit software simulation tool called PSICE but more insight into the operation of the device was obtained by using a concept from the nonlinear dynamics community, the Liénard oscillator [7]. The details of this Liénard oscillator model are covered in chapters 6 and 7 but in broad outline the Liénard oscillator is a generalisation of the more well-known van der Pol oscillator model which was first used in the 1920s to describe

devices and systems that exhibited negative differential resistance. The van der Pol oscillator uses a third order polynomial to represent the I – V curve of an electronic oscillator—it is the lowest order polynomial that has an NDC feature and so the van der Pol oscillator is the simplest model of an NDC oscillator [8]. However, the Liénard oscillator model uses higher order polynomials which can more accurately represent the NDC curve of any given device or system. For our application, the RTD—LD hybrid circuit, the Liénard oscillator resulted in a more accurate model of the system and allowed the oscillation output of the RTD—LD to be closely fitted as is shown next.

The details of the model are given in [7]. A schematic of the physical layout of the circuit is shown in figure 5.3. By applying Kirchhoff's laws and the usual AC circuit analysis of the circuit shown in figure 5.4, we obtain equations (5.1) and (5.2); they can be used to derive the time dependence of the I – V curve of the RTD:

$$\frac{dV(t)}{dt} = \frac{I(t) - f(V)}{C} \quad (5.1)$$

$$\frac{dI(t)}{dt} = \frac{V_b - RI(t) - V(t)}{L} \quad (5.2)$$

where $V_b = V_{DC} - V_{th}$, and V_{DC} is the DC bias voltage, $I(t)$ is the current through R and L and $V(t)$ is the voltage across the RTD; $f(V)$ is the highly nonlinear voltage–current characteristic of the RTD and is discussed in chapter 2. Because of the NDC

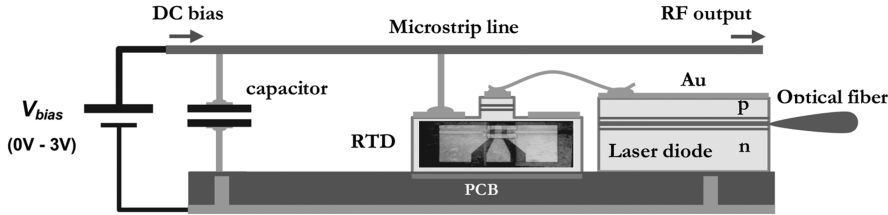


Figure 5.3. The figure represents the electrical connections on a hybrid RTD—LD printed circuit board.

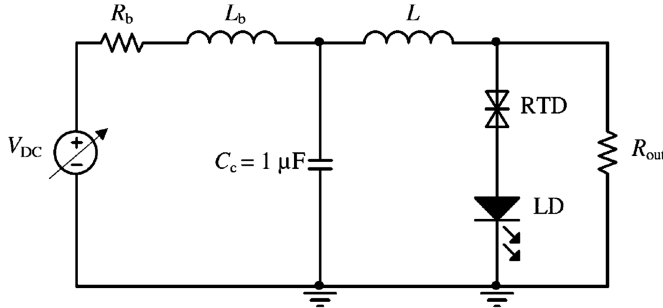


Figure 5.4. Schematic of the hybrid RTD—LD electrical circuit, R_b , is the resistance associated with the bias circuit L_b , is the bias circuit inductance, L is the value of the inductance of the wire connecting the RTD and the LD C_c is the capacitance, R_{out} is the output resistance.

associated with $f(V)$, the circuit shown in figure 5.4 oscillates—the oscillation frequency is set by the values of the external components in the circuit.

The electrical current from the RTD is injected into the laser and in the following equations N is the electron density and S is the photon density:

$$\frac{dN}{dt} = \frac{I(t)}{q\mathcal{V}} - \frac{N}{\tau} - g_0(N - N_0) \frac{S}{1 + \epsilon S} \quad (5.3)$$

$$\frac{dS}{dt} = g_0(N - N_0) \frac{S}{1 + \epsilon S} - \frac{S}{\tau_p} + \frac{\beta N}{\tau} \quad (5.4)$$

$I(t)$ is the current through the laser equation (5.2); $q = |1.6 \times 10^{-19} \text{ C}|$ is the electronic charge, τ is the lifetime of an electron in the active mode volume, $\mathcal{V}$, τ_p is the lifetime of a photon in the active mode volume, N_0 is the minimum electron density required to obtain net gain, g_0 is the gain coefficient, β is the spontaneous emission factor, and ϵ is the nonlinear gain compression factor. Equations (5.3) and (5.4) describe how the electrical current from the RTD, $I(t)$ is injected into the active region of the laser diode and converted into an optical output.

Solutions to equations (5.2)–(5.4) are used to predict the optical power output from the RTD—LD. The comparison of the model to the experimental data is given in figure 5.5.

The Liénard oscillator model also predicts that the RTD—LD will operate as an optoelectronic, voltage controlled oscillator (VCO) [7]. That is, the fundamental oscillation frequency will depend on the bias voltage, V_{DC} , applied to the RTD—LD. Figure 5.6 shows the VCO characteristics of the RTD—LD and the fit of the Liénard oscillator VCO model to the experimental data of the DC bias voltage

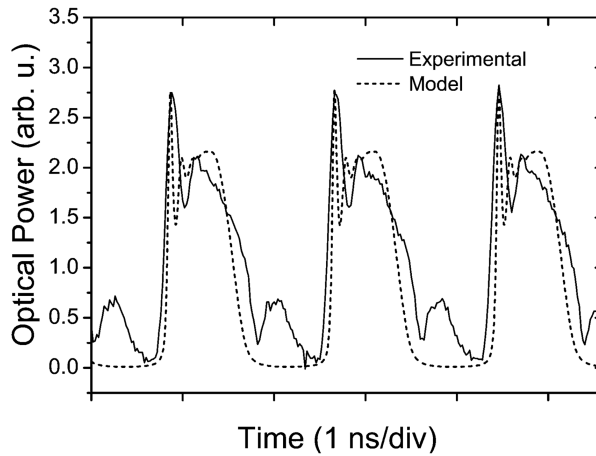


Figure 5.5. Experimentally measured output from the hybrid RTD—LD (solid line) modelled (dashed line) using the Liénard oscillator model; the DC bias was set to 1.8 V and the fundamental frequency of oscillation is 560 MHz [7].

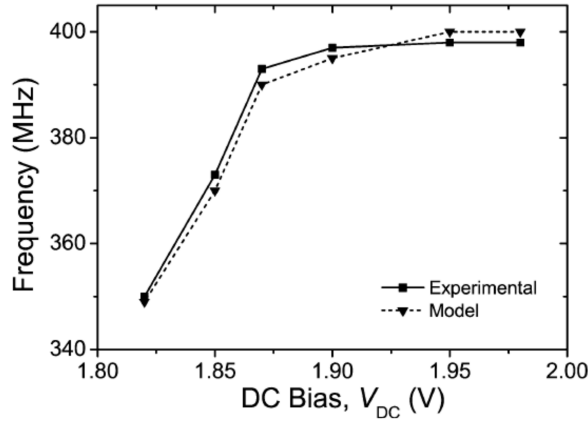


Figure 5.6. The fundamental frequency of the RTD—LD device versus the applied DC bias—the solid line is the experimental data and the dashed line is the curve fitted using the Liénard oscillator model.

dependent frequency of the RTD—LD. A low inductance version of the same device could be tuned 1.8–2.15 GHz by adjusting the DC bias—see also [7].

The nonlinear model of the RTD—LD oscillator is more complete than linear models. The linear models of oscillator circuits [9] are reasonably accurate in predicting oscillation frequencies but often fail to predict output powers or oscillation waveforms.

5.3 Integrated RTD—laser circuit

Previous work on monolithic integration of RTDs and AlGaAs/GaAs laser diodes had been on laser emitting at ~850 nm [3]. There has been also work on using a multiple quantum well region as part of an RTD to obtain NDC and bistable operation [10]. Further, there had been some work on observing the resonant tunnelling into quantum wells in a multiple quantum well laser but without the explicit addition of a double barrier RTD [11]. Tunnelling has been used to achieve large modulation bandwidths, 98 GHz, in laser diodes operating at 980 nm [12]. For a theory treatment of the properties of monolithic integration of a resonant tunnelling diode and a quantum well laser—see [13]. In this section, we discuss the first RTD monolithic integration with a laser diode emitting at optical communication wavelengths (around 1500 nm). The epitaxial layer, waveguide configuration and equivalent electrical circuit of an integrated RTD—LD designed to operate at optical communication wavelengths [2, 14] is shown in figure 5.7.

The RTD consisted of three epitaxial layers: a 5 nm InGaAs quantum well between two strained (lattice mismatch) 2 nm AlAs barriers. The laser section consisted of six 6.7 nm thick InGaAs quantum wells with 10 nm InGaAlAs barriers. Except for the AlAs RTD barrier layers, all epitaxial layers were lattice match to InP which was used as the substrate. The AlAs layers are thin enough that they do not disrupt the crystal growth because they are smaller than the critical layer thickness [15].

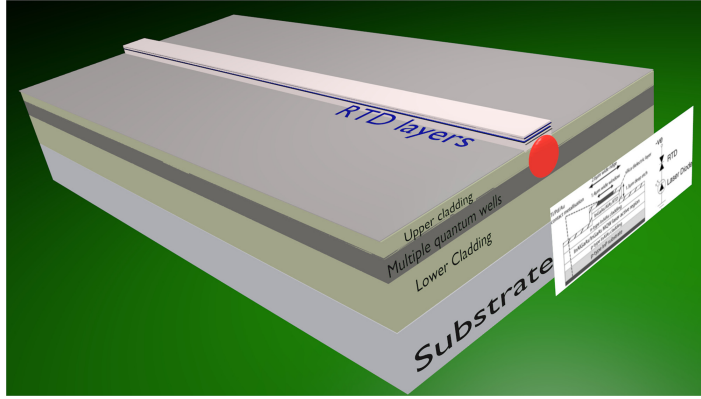


Figure 5.7. Schematic of RTD—LD chip—it is a metal clad ridge waveguide laser with the RTD in the mesa that forms the ridge; the red dot indicates the region of the chip facet where light is emitted—a cross section view is shown as an inset—for a better view of the cross section, see figure 5.8.

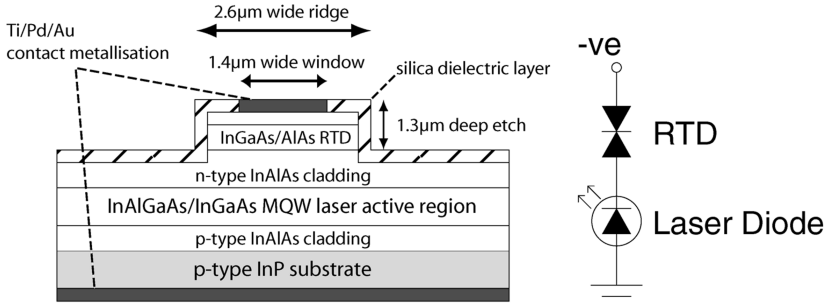


Figure 5.8. Cross section of the layout of the integrated ridge waveguide RTD—laser diode (RTD—LD). Laser resonator length is $500\text{ }\mu\text{m}$ giving an RTD active area of $1300\text{ }\mu\text{m}^2$; on the right is a schematic lumped element circuit representation of the device [2].

The RTD section was embedded in *n*-type material because the RTDs have conventionally operated with electrons rather than holes, plus it was necessary to place the RTD in the waveguide mesa so that the current density through the device was sufficiently large to reach the NDR region of the RTD I – V curve with currents low enough not to cause thermal damage (figure 5.8). Usually semiconductor lasers are ‘*p*-side up’, that is, the waveguide channel layers furthest from the substrate are *p*-type doped. However, for the RTD—LD with RTD layers on top of the waveguide in *n*-type material means that the RTD—LD is *p*-side down.

At room temperature this RTD—LD operated at 1540 nm , but the device could only be operated in pulse mode. To explore the continuous wave operation of the device and to clearly demonstrate the bistable operation it was necessary to operate at cryogenic temperatures (130 K).

Figure 5.9 clearly shows that the device was bistable at cryogenic temperatures, that is, the device has two stable states in terms of the DC bias applied as the DC bias is increased the device output flips on and then as it is decreased the device flips

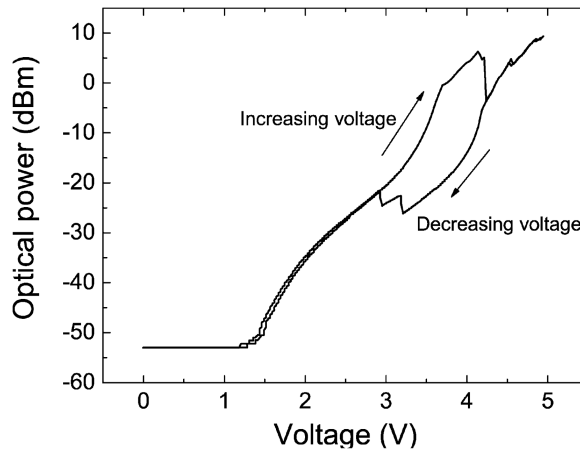


Figure 5.9. The optical output (in dBm) from the integrated RTD—LD operating 130 K versus the DC bias voltage—the device shows clear bistability and hysteresis with >30 dB between the on and off states.

off at a lower voltage than the switch-on voltage that occurs as the voltage was increased. This means the device has a hysteresis type of operation where the current state depends on the history of past operation. The difference between the on and off optical power output was ~30 dB.

The monolithic integration of a double barrier RTD and laser diode has been demonstrated at optical communications wavelengths and the RTD—LD device shows bistable operation which could be particularly convenient for digital switching and optical memory operations [10].

5.4 Conclusions

The RTD can be integrated with laser diodes to form RTD—LD devices. Two integration layouts have been discussed: (i) the hybrid integration where there are two separate chips, the RTD and the laser diode chips, that are connected with a short wire (~2 mm), and (ii) the fully monolithic integrated layout with the RTD forming part of the epitaxial layers that make up the top contact layers in a semiconductor laser chip.

When DC biased in the NDC region the hybrid RTD—LD optical power output oscillates at frequencies set by the value of the external components—these frequencies can be >1 GHz.

Although so far operating CW only at cryogenic temperatures (~130 K), the monolithically integrated RTD—LD has emitted at wavelengths (1540 nm) in the C band of the optical communications window and has demonstrated bistable operation [2].

In chapters 6 and 7, we discuss how the large electrical gain of the RTD—PD (chapter 4) can produce enough electrical power to drive the hybrid RTD—LD to form part of an optoelectronic oscillator (OEO) that employs optical fibre as the energy storage part of the oscillator. The RTD-based OEO is a version of the OEO

that employs a minimal number of components in a highly efficient, low noise, high Q oscillator operating at rf frequencies >1 GHz.

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Resonant Tunneling Diode Photonics

Devices and applications

Charlie Ironside, Bruno Romeira and José Figueiredo

Chapter 6

Nonlinear dynamics of RTD oscillators

6.1 Introduction

There is substantial interest in the development of novel transmitters and receivers to meet the future needs of communication and computing systems. Among various approaches, the implementation of electro-optical and all-optical solutions that take advantage of the unique nonlinearities of semiconductor active devices (e.g. optoelectronic oscillators, laser diodes and modulators) have attracted considerable attention in the last decades. The key applications of nonlinear photonic devices include microwave-photonic systems for signal processing [1, 2], secure optical chaos communications [3, 4], and more recently neuromorphic (brain-inspired) photonic applications [5].

As was thoroughly discussed in chapters 5 and 4, the recent advances on RTD—LD and RTD—PD modules, respectively, established a new-type of RTD-based photonic circuit that led to significant breakthroughs including the demonstration of various dynamical regimes such as voltage-controlled autonomous relaxation oscillations in both electrical and optical domains. In this chapter, we cover the nonlinear dynamic properties of non-autonomous RTD photonic oscillators, that is, subjected to external perturbations. The key dynamic scenarios covered in this chapter include injection locking, period-adding bifurcation, quasi-periodic route to chaos, and excitable (i.e. neuron-like) dynamics.

6.2 Photonic synchronisation and chaos

6.2.1 Injection locking

Injection locking is a well-known phenomenon discussed in standard textbooks on nonlinear oscillators [6, 7], under one of various headings such as synchronisation, phase locking, frequency entrainment, or forced oscillation. According to the oscillator model developed by Adler and described in the seminar paper [8], an oscillator can be locked in frequency by an external signal which is injected into the

oscillator and the mechanism of the locking process depends upon the initial frequency difference between the oscillator, the relative amplitude between the injected signal and the oscillator, and the respective circuit parameters.

In this section, we investigate the injection locking characteristics of RTD—LD oscillators to incoming RF signals. In figure 6.1(a), a photograph of a typical RTD—LD hybrid circuit module mounted on a printed circuit board is shown (see chapter 5 for more details on the RTD—LD circuit and its free-running autonomous operation). The respective diagram showing both the equivalent circuit and the experimental setup is depicted in figure 6.1(b). For the injection locking experiment, an electrical signal is injected into the electrical input port using a bias-T and the output signal is coupled from the LD optical output using a lensed fibre to a photodetector. The RF spectra of the free-running and injection-locked RTD—LD outputs are typically measured using a fast spectrum analyser or oscilloscope.

In figure 6.2(a) a typical spectrum (black dashed-dot) of RTD—LD's free-running oscillations is presented (in this example, the oscillation frequency is set to around 1.24 GHz). The corresponding single side band (SSB) phase noise measurements are shown in figure 6.2(b). The photo-detected free-running signal from the laser output shows a low-quality phase noise and a large frequency drift and instability. The corresponding single-sided noise (SSB) power density is -81.96 dBc/Hz at 100 kHz offset of the carrier centre frequency.

When an RF signal frequency is injected at a frequency sufficiently close to the free-running oscillation frequency (even using extremely low power levels, in this case -43 dBm), the RTD electronic oscillation is stabilised and pulled to the same frequency of the incoming RF signal. The corresponding modulated laser output signal is also injection locked and highly-stable with a corresponding SSB phase noise of -85.86 dBc/Hz noise at 100 kHz offset of the centre frequency. We note that, in the case of the injected power being lower than the free-running signal, the amplitude of the injection-locked RTD—LD is similar to the amplitude of the free-running signal. On top of figure 7.2(a) is an inset with the spectrum surveillance confirming the spectral stability of the locked signal. The pull or locking range Δf was also measured. For this level of injected power, the locking range was around

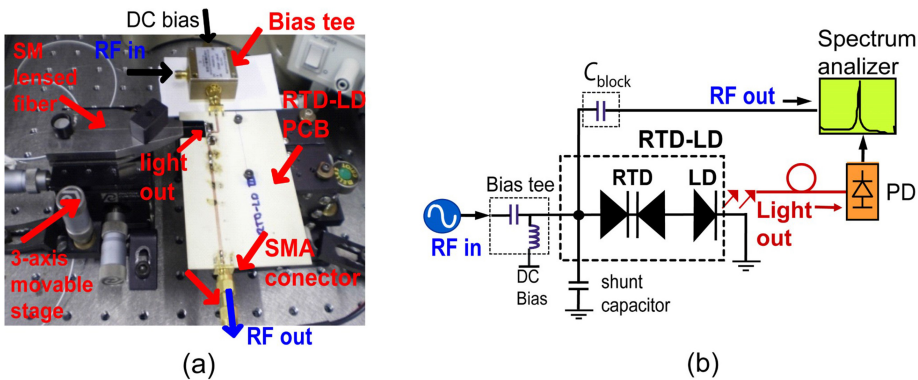


Figure 6.1. (a) Photograph of RTD—LD circuit. (b) Experimental setup and circuit schematic diagram.

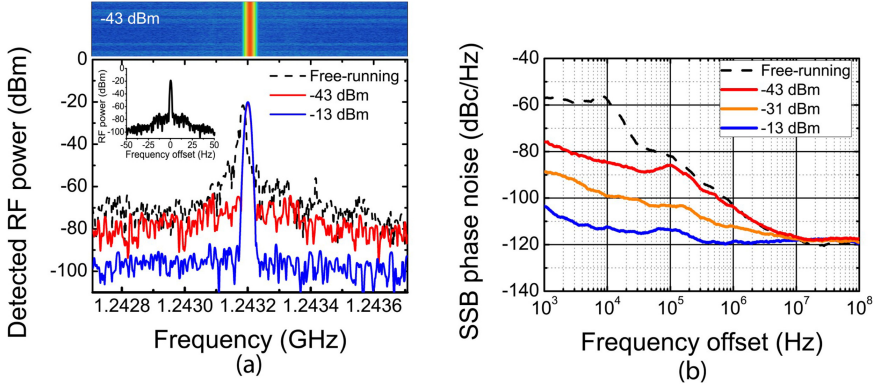


Figure 6.2. (a) Free-running and injection-locked RF spectra for the photo-detected RTD—LD outputs (1 MHz span and 10 kHz RBW). The inset shows the injection-locked spectrum to an incoming RF signal at -13 dBm power level (100 Hz span and 1 Hz RBW). Also shown at the top is a spectrogram that represents the evolution of spectral density of the injection-locked signal recorded over approximately 60 s (the vertical axis represents time). (b) SSB phase noise measurements for both free-running oscillations and injection-locked signals.

300 kHz. For injected power levels much lower than the power of the RTD—LD free-running oscillator, the locking range was found to be approximately proportional to $\sqrt{P_i/P_0}$, where P_i is the injected power and P_0 is the RF output power of the RTD—LD free-running signal [9]. This behaviour follows Adler's theory [8], and the injection locking mechanisms found in other conventional negative-resistance oscillators operating in the small-signal limit.

We note that increasing the output power of the RF synthesiser further decreases the phase noise level towards the reference signal. As the noise level of the input is decreased, the phase noise of the injection-locked signal decreases accordingly. This is demonstrated in the spectrum (blue trace) of figure 6.2(a) using a power level of the incoming RF signal of -13 dBm. Figure 6.2(b) shows the corresponding low phase noise with -113.69 dBc/Hz phase noise at 100 kHz carrier offset. The injection-locked RTD—LD oscillator follows the noise characteristics of the input signal, limited only by the measurement instrument noise floor. The inset of figure 6.2(a) shows the same injection-locked signal but using a 1 Hz resolution bandwidth (note the different frequency scales), confirming that the output spectrum collapses to match that of the reference synthesiser. We achieve a suppression of more than 30 dB when compared with the free-running oscillating signal. The corresponding locking range increased to 30.5 MHz, i.e. approximately 2.45% of the oscillation frequency.

6.2.2 Period-adding bifurcation

In order to map the dynamics of the RTD—LD photonic circuit subjected to external injection, we have characterised in detail the electro-photonic outputs in both time and Fourier domains for a wide range of voltage amplitude V_{AC} and injected frequency f_{in} values. The observed dynamic trajectories were plotted in the

voltage-light (V – S) phase space, where V represents the voltage across the RTD—LD, the electrical output, and S is the LD optical output. In the free-running case, the trajectories in the V – S phase space correspond to a simple attractor, also known as a limit cycle, shown in figure 6.3(a). The corresponding free-running frequency was $f_0 \sim 0.95$ GHz. We note in this example the circuit analysed was oscillating at frequencies, f_0 , ranging from 0.5 GHz to ~ 1 GHz, controlled by the DC voltage [10]. The corresponding spectra of both the free-running electrical and optically modulated oscillations are presented in figure 6.3(b) showing a harmonic content up to the tenth harmonic.

Following the injection locking phenomena discussed in the previous subsection, a more complex circuit behaviour known as period-adding bifurcation dynamics [11] can be obtained when the free-running oscillations are perturbed by injected RF signals at frequencies that do not correspond to the fundamental frequency of the oscillator. In order to map these dynamic regimes, whenever a stable closed trajectory was found in the V – S phase space the system, we considered this a phase-locked regime. We also considered a very weak perturbation regime, $V_{AC} \ll \Delta V_{NDC}$, where ΔV_{NDC} is the NDC voltage width (~ 250 mV for the circuits tested). In these conditions, the free-running and injected frequencies, f_0 and f_{in} , satisfy the relation $f_{in}/f_0 \rightarrow n/m$, where n and m are natural numbers. The ratios n/m are known as winding numbers [6, 7]

Figures 6.3(c)–(j) present examples of RTD—LD synchronisation states for sinusoidal input amplitude voltage of 300 mV and scanning the injected frequency from 0.3 GHz to 3.5 GHz. On the left are shown the V – S phase space diagrams and on the right the corresponding spectra. Figures 6.3(c)–(d) show locking at a frequency which is half of the free-running oscillation (1:2). Figures 6.3(e)–(f) show locking at signals with a frequency close to the free-running oscillation (1:1)—fundamental locking; figures 6.3(g)–(h) and figures 6.1(i)–(j) illustrate two examples of harmonic injection locking, 2:1 and 3:1, respectively, where the injected frequencies are close to RTD—LD first and second harmonic frequencies, respectively.

The overall circuit dynamic response to perturbations was mapped in a two-dimensional Arnold tongues diagram and is shown in figure 6.4(a). The tongues represented in the map correspond to the regions of locking to the driving signals presented in figures 6.3(c)–(j). The Liénard’s numerical model, see section 6.3, agrees qualitatively well with experimental tongues. The y -axis represents the perturbation strength defined as the ratio of the injected voltage amplitude V_{AC} and RTD—LD peak voltage V_P (around 2.90 V), V_{AC}/V_P ; the x -axis corresponds to the frequency ratio f_{in}/f_0 . The injected amplitude and frequency ranged from 50 mV to 500 mV and 0.3 GHz to 3.5 GHz, respectively. Figure 6.4(b) presents the corresponding simulated bifurcation map of the photon density, $s(t)$, showing the period-adding bifurcation dynamics sequence between intervals of unlocked signals. The results presented in figure 6.4 demonstrate the RTD—LD circuit following a period-adding bifurcation dynamics. The modelling results confirm that under the regimes of operation analysed here the RTD—LD circuits follows RTD’s nonlinearity in a way that the laser dynamics follows exactly the dynamics of the RTD oscillator.

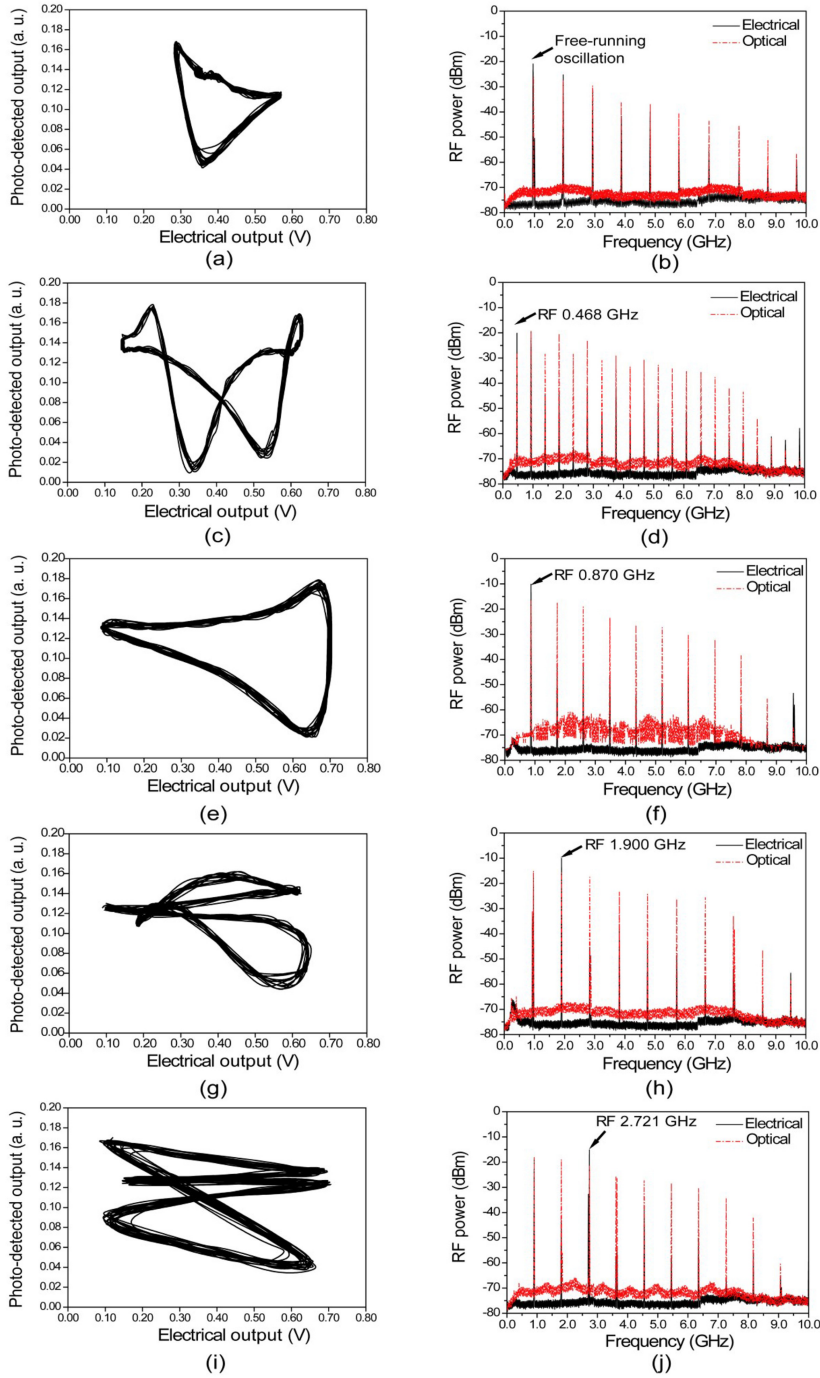


Figure 6.3. Experimental electro-phonic limit cycle regimes in the V - S phase space (left) and corresponding spectra (right) for the following stable dynamic states: (a), (b) free-running; and locked (c), (d) 1:2 and (e), (f) 1:1, (g), (h) 2:1, and (i), (j) 3:1.

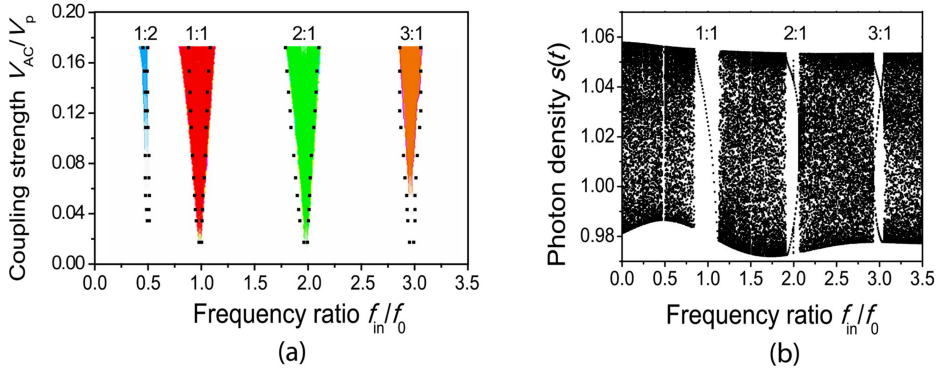


Figure 6.4. (a) RTD—LD oscillator frequency-locking structure showing the electro-photonics Arnold tongues. The results from the experiment (dotted lines) are compared with theory (coloured tongues). (b) Numerical simulated bifurcation map of photon density $s(t)$ as a function of frequency for electrical injection of a periodic signal for a fixed low amplitude.

6.2.3 Quasi-periodic route to chaos

Outside the phase-locked regions the circuit provides unsynchronised complex outputs that can be regarded as a result of the nonlinear mixing of the injected signal with the free-running oscillations. This results in quasi-periodic signals displaying a large frequency content, and chaotic signals with a broadband (GHz wide) spectrum. Figure 6.5(a) shows an example of a quasi-periodic trajectory in the unsynchronised region between the winding numbers 1:1 and 2:1. Contrary to the previously synchronised orbits, quasi-periodic signals trajectories do not form a closed loop. Instead, the orbit circulates around a confined region designed as quasi-periodic torus. Figure 6.5(b) shows the corresponding power spectra of V and S , showing a spike-like frequency distribution, both with the same frequency content. Increasing the nonlinearity by injecting a signal with amplitudes much larger than 300 mV the circuit dynamics evolves from synchronisation to quasi-periodicity and then onto chaos, where chaotic burst signals start to appear outside the torus as the frequency ratio f_{in}/f_0 converge to a certain irrational number. We observed experimentally these complex dynamics in a wide range of injected amplitudes. Figure 6.5(c) shows a typical chaotic trajectory for a driving amplitude of 790 mV. The corresponding spectra of the chaotic signal is displayed in figure 6.5(d). The sharp spike-like frequency peaks of quasi-periodic signals, figure 6.5(b), disappear in the transition to the chaotic regime, giving rise to signals with a broadband continuous frequency distribution. The peaks observed in figure 6.5(d) correspond to the injected RF signal and to minor vestiges of some quasi-periodicity, an indication that the system's dynamics follows a quasi-periodic route to chaos. Also observed is a substantial increase of the noise level (up to 20 dB), a key characteristic of chaotic signals.

The intrinsic instability and sensitivity to initial conditions of the chaotic signals in the RTD—LD photonic circuits was numerically investigated by evaluating the rate of divergence of nearby trajectories in the unsynchronised regions. Figure 6.6

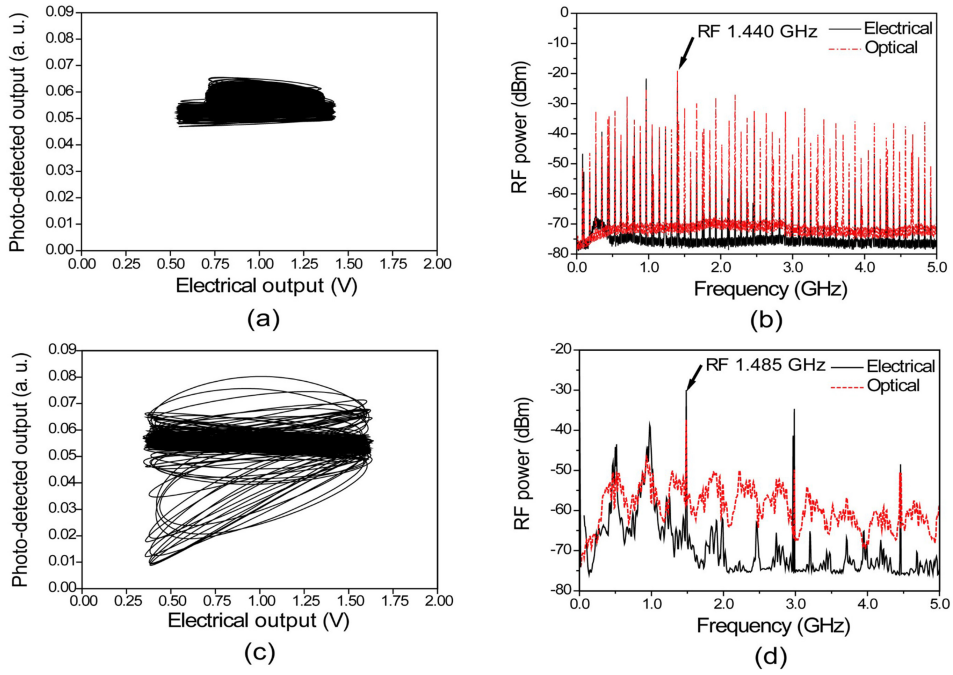


Figure 6.5. Experimental electro-phonic regimes in V – S phase space (left) and corresponding power spectra (right) for (a), (b) quasi-periodic, and (c), (d) chaotic dynamic states.

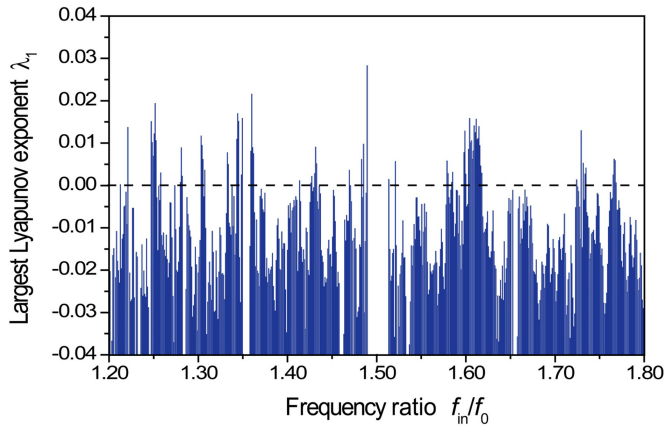


Figure 6.6. Histogram of the largest Lyapunov exponent for the unsynchronised region between 1:1 and 2:1 locking windows for the experimental conditions presented in figure 6.5.

presents a histogram of the largest Lyapunov exponent λ_1 , illustrating the frequency bands where the RTD–LD behaves chaotically, that is, where $\lambda_1 > 0$. The largest chaotic band corresponds to f_{in}/f_0 around the golden number $(\sqrt{5} + 1)/2 \sim 1.62$ (also called the golden mean). The golden number, and other irrational numbers including the silver number, are commonly used in nonlinear dynamical studies [6]. The

frequency range of chaotic operation and the chaos signatures, namely the broadband spectrum shown in figure 6.5(d), indicate the circuit is operating in the golden ratio band [6] the largest chaotic band shown in figure 6.6, since the intrinsic free-running frequency red shifts with the amplitude of the injected signal, leading to a value of f_{in}/f_0 close to the golden ratio. The analysis of figure 6.5(d) indicates the free-running frequency (second peak in the spectrum) under this level of injection was shifted from 953 MHz to ~ 930 MHz, which gives $f_{\text{in}}/f_0 \sim 1.6$.

As a summary, the process of achieving broadband chaotic optical sub-carriers in the laser output described here is physically different from the methods presented in [4, 12, 13]. The generation and conversion of chaotic currents into chaotic light intensity patterns discussed here depend solely on the RTD nonlinearities. Since monolithic photonic integration leads not only to a considerable reduction in size, low energy consumption and reduced cost but also to an increase in speed and reliability, RTD—LD monolithic photonic circuits have potential as an alternative to the chaotic systems reported, for example in [13] to operate as the core of electro-photonics chaotic broadband emitters to encode data at the physical level.

6.3 Excitable spiking in neuromorphic photonic devices and systems

6.3.1 Excitability (neuron-like) dynamics

Excitability is a well-known nonlinear property [14] that regulates several biological process like nervous impulse propagation [15] or chemical reactions such as Belousov–Zhabotinsky systems [16]. Excitable systems provide an all-or-none response to external stimuli: the system responds linearly and with small amplitude to stimuli that do not overcome a threshold, but otherwise it yields a large pulse (whose shape is independent of the stimulus) followed by a lapse, called the refractory time, during which the system does not respond to new stimuli. Excitable systems show potential to perform complex information processing tasks for applications in clock recovery and pulse reshaping functionalities in photonics [17]. However, the currently proposed systems operate at low speed and are bulky making most of them not suitable for present day information processing needs.

In this section, we investigate a high-speed neuromorphic photonic circuit using the RTD—LD photonic circuit configuration analysed in the previous sections. The typical circuit used to investigate the excitable dynamics is shown in figure 6.7.

The circuit displays an asymmetric DC current–voltage (I – V) curve with a pronounced negative differential conductance region, figure 6.7(d). When DC biased in the NDC, they can operate as high-speed oscillators for THz generation [18] and recently RTD-based OEOs were proposed for high-purity phase-locked oscillations employing either external injection locking, as analysed in the previous section of this chapter, or self-injection locking in a delay line feedback configuration [19] as analysed in chapter 7.

However, when DC biased outside the NDC, RTD-based photonic circuits can display new phenomena including (i) bistable behaviour [20] and (ii) excitable behaviour as shown recently in [21] where optical pulse generation with lapse time responses of the order of nanoseconds was demonstrated. In this section,

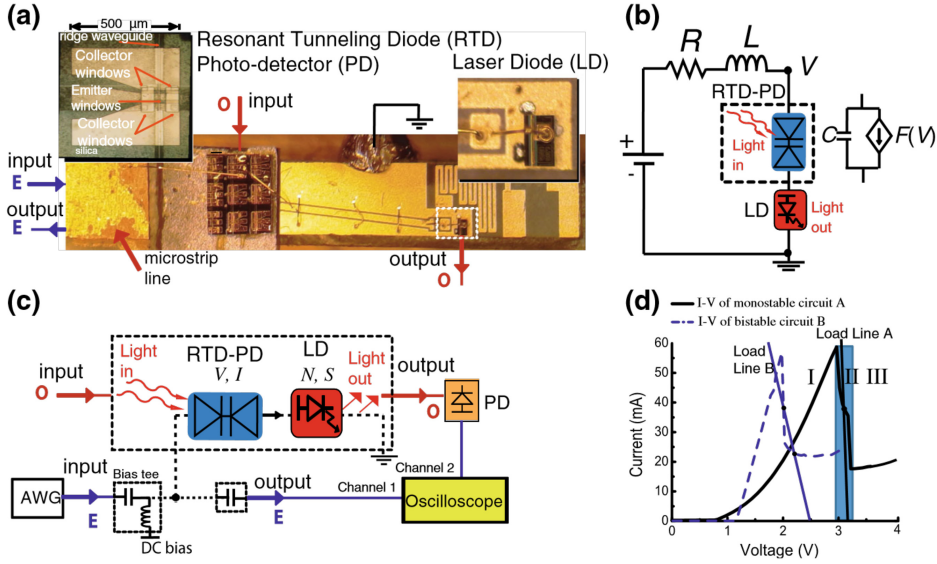


Figure 6.7. (a) Photograph of the hybrid optoelectronic integrated circuit. Inset shows the RTD and LD dies. (b) Corresponding lumped electrical circuit showing the equivalent inductance L , resistance R , and RTD—PD capacitance C and I - V nonlinear $F(V)$. (c) Schematic block diagram of the RTD in series with the LD forming the optoelectronic oscillator with quadruple electronic and optical inputs/outputs. (d) Experimental I - V curves of the monostable (solid) and bistable (dashed) circuits.

we cover this excitability phenomena characterised by noise activated dynamics when RTD-based circuits operate in either monostable or bistable regimes. In these regimes, we observe spikes, square pulses, bursting, and stochastic and coherence resonances depending on the working conditions of the RTD photonic circuits. The experimental results are well described employing a nonlinear Liénard oscillator model—that describes the RTD stochastically activated by means of white Gaussian noise—driving the LD, described by single mode rate equation, see chapter 5. The stochastically induced dynamics can be exploited in a variety of applications including optoelectronic memories, or in artificial neural networks by coupling multiple excitable units.

6.3.2 Spiking and bursting dynamics

Figure 6.7(a) shows a photograph of the experimental circuit, and figure 6.7(b) presents the equivalent electrical circuit. We measured the response of the circuit to white Gaussian noise signals injected using an Agilent 33250A arbitrary waveform generator (AWG), by recording the electrical and optical outputs employing a 2 GHz oscilloscope, figure 6.7(c).

The operating quiescent point of our circuit is given by the intersection of the nonlinear I - V curve and a load line whose slope is defined by the inverse of the resistance of the circuit. As shown in figure 6.7(d), for low resistance the load line is almost vertical and there is only one intersection with the nonlinear characteristic

function that enables operation in either a rest state regime, regions I and III, or a self-oscillatory regime, region II, depending on the bias conditions, while under high resistance conditions several (three) intersections may occur.

The equivalent photonic circuit is described by single mode laser rate equations describing LD's photon and carrier density (S , N) coupled to a Liénard system which accounts for the current and voltage across the RTD—LD, as described in chapter 5 and summarised here as follows:

$$\dot{V} = \frac{1}{\mu}[I - f(V) - \chi\xi(t)] \quad (6.1)$$

$$\dot{I} = \mu[V_0 - \gamma I - V] \quad (6.2)$$

$$\dot{N} = \frac{1}{\tau_n} \left[\frac{1}{I_{th}} - N - \frac{N - \delta}{1 - \delta} (1 - \epsilon S) S \right] \quad (6.3)$$

$$\dot{S} = \frac{1}{\tau_p} \left[\frac{N - \delta}{1 - \delta} (1 - \epsilon S) S - S + N\beta \right] \quad (6.4)$$

Here, time has been scaled by the LC resonance frequency $\omega_0 = (\sqrt{LC})^{-1}$, $f(V)$ describes the I – V curve, and $\mu = V_0/I_0\sqrt{C/L}$ and $\gamma = R(I_0/V_0)$ describe the equivalent circuit parameters (inductance, L , capacitance C , and resistance R , see figure 6.7(b)). Finally, we introduce the stochastic effects in our model employing an effective delta-correlated Gaussian white noise $\xi(t)$ of zero mean and unit variance, so χ stands for the total noise strength.

When the RTD—LD operates in monostable conditions and is DC biased within the NDC, region II, figure 6.7(d), it exhibits voltage-controlled self-sustained current oscillations driving the laser. When DC biased in regions I and III, figure 6.7(d), the OEO responds to external perturbations by generating spikes in both electrical and optical domains when the perturbation exceeds a given threshold [21]. In this excitable regime, the system is highly sensitive to noise emitting fixed-form spikes or bursts of spikes depending on the DC bias point, and showing coherence resonance (CR) [14], i.e., a maximum of temporal regularity in the output for a finite noise level. Figure 6.8 shows a numerical example of the occurrence of spiking and CR behaviour in the RTD—LD due to noise perturbation, and selecting the bias operation in region I, i.e., in an equilibrium situation in the positive differential resistance. For a very weak noise input level, $\chi = 0.12$, the RTD—LD output maintains its steady (rest) state. For a given noise threshold perturbation, however, the RTD—LD can leave the rest state. When $\chi = 0.13$ (weak noise) we observe the output spiking in both electrical and laser outputs (V and S , respectively). The peaks appear random, and the interval between excitations varies substantially. For a moderate noise, $\chi = 0.18$, the spiking is rather regular, which implies that the inter-spike intervals do not differ much, providing an example of CR behaviour.

The pulse triggering mechanism is explained as follows. The RTD—LD can be considered as a slow–fast excitable system in which the fast variable is the voltage

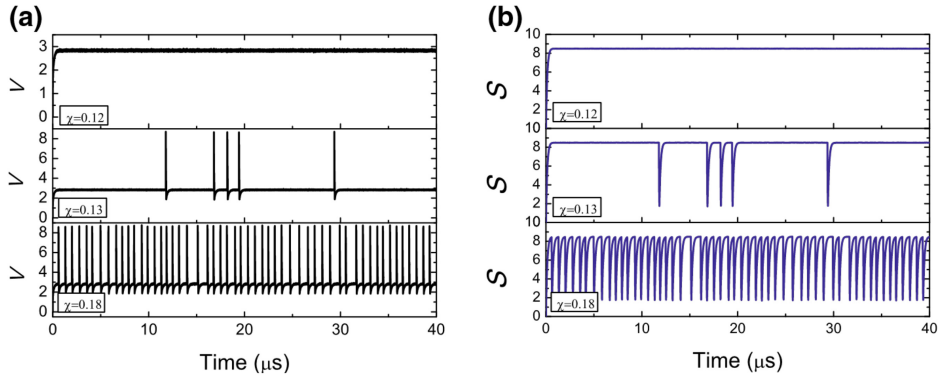


Figure 6.8. Simulated noise induced resonances in the RTD—PD—LD excitable system in both (a) electrical (V) and (b) optical (S) domains as a function of noise strength.

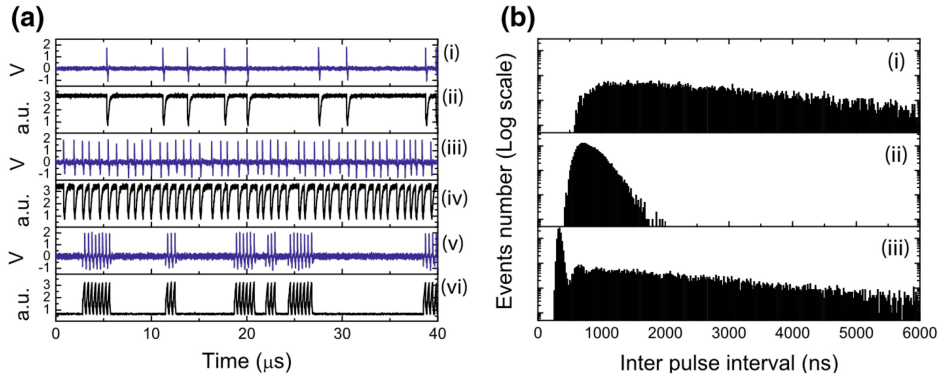


Figure 6.9. (a) Experimental time traces of electrically noise induced neuron-like phenomena in both the electrical (V) and the optical (a.u.) domains: (i)–(ii) random pulsing firing ($V_{DC} = 2.85$ V and $V_{noise} = 180$ mV); (iii)–(iv) coherence resonance behaviour ($V_{DC} = 2.85$ V and $V_{noise} = 225$ mV); (v)–(vi) multi-pulsing bursting behaviour ($V_{DC} = 3.36$ V and $V_{noise} = 250$ mV). (b) Corresponding histograms of the ISI statistics of the laser output: (i) random pulsing firing; (ii) coherence resonance behaviour; (iii) multi-pulsing bursting behaviour.

and the slow one the current. During the excitable orbit, the two fast stages correspond to sudden rise and fall of the voltage during which the current does not change appreciably. These two fast stages are interleaved by two slow stages in which the current evolves along the attracting nullcline as defined by the $f(V)$ function while the voltage follows adiabatically.

The excitable regimes predicted in the numerical simulations were verified experimentally. By adding to the DC bias in region I a stochastic voltage generated by a Gaussian noise source with a cut-off frequency of 80 MHz, the RTD—LD circuit A passes from emitting occasional spikes (see figure 6.9(a), plots (i) and (ii)) to emission of an almost periodic pulse train (see figure 6.9(a), plot (iii) and (iv)) as the noise level increases. The upward electrical pulses have a full width half maximum (FWHM) of ~ 13 ns, while the LD light output exhibits downward pulses of FWHM of ~ 200 ns. The LD emits pulses in response to current modulation and its response

is determined by carrier and photon lifetimes, yielding optical pulses longer than the current spikes. The photonic circuit has potential to achieve sub-ns response [21], limited by the parasitics, e.g. the gold wire's inductance, and by the laser relaxation oscillation. Further increasing the noise level above 225 mV destroys the periodicity of the signal, yet the pulse shapes remain. Further analysis of the corresponding stochastic induced dynamics inter-spike interval (ISI) series data, figure 6.9(b), reveals the pulse statistics showing a typical exponential behaviour of a Kramer's escape process [14], displaced by the refractory time of the excitable orbits. Figure 6.9(b) (ii) shows the ISI for the case of coherence resonance.

Similar behaviour is observed when the RTD—LD is biased in region III, but now emitting downward—upward electrical pulses, figure 6.9(a), plot (v), and upward optical pulses that appear in bursts (figure 6.9 plot (vi)) due to the asymmetry of the I – V curve, which has a smaller slope on the valley side. Note that, since the operating point is now at the low current value in region III, which defines the starting point of the excitable orbit, the system experiences a current increase which explains the upward optical pulses. The analysis of the inter-spike interval (figure 6.9(b) (iii)) reveals that the bursts manifest themselves in the histograms as a peaked structure at n/T_1 , where $T_1 = 350$ ns corresponds to the excitable refractory time. This means that in this bursting regime, if we have an excitable orbit at T_1 , it is likely to be followed by another one at $T_2 = T_1 + T_1$. We see also the appearance of the statistics with its normal tail, that is very similar to the figure 6.9(b) (i), as a result of the excitable response.

6.3.3 Bistable dynamics

When the RTD—LD is loaded with high resistance (the circuit shown in figure 6.7 (b)), it shows a region of bistability around 2 V, figure 6.7(d). This bistability is evidenced as noise-triggered hopping between the two stable states in regions I and III. When DC biased in such a condition, switching between these two states leads to some irregular waveforms composed of two plateaux whose duration is governed by the Kramer escape time in both basins of attraction, figure 6.10, plots (i) and (ii). Increasing the noise level reduces the residence times in both states, and leads to an almost regular square waveform, see figure 6.10(iii). This is the basis for stochastic

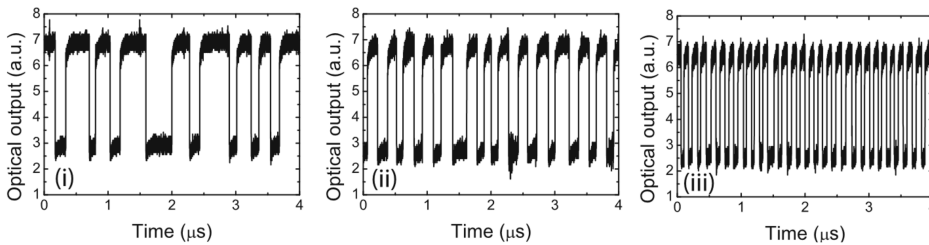


Figure 6.10. Experimental time traces of electrically noise induced neuron-like phenomena measured in the laser output for a circuit DC biased in a bistable condition, $V_{DC} = 2$ V. Random fired squared pulses at (i) $V_{noise} = 125$ mV, and (ii) $V_{noise} = 150$ mV; (iii) periodic sequence of square pulses, $V_{noise} = 250$ mV.

resonance, whereby a signal of small amplitude can be amplified and regenerated in bistable systems by adding an optimum amount of noise.

6.4 Conclusions

In this chapter, we have covered the nonlinear dynamics of semiconductor devices and systems based on resonant tunnelling. Fundamentally the resonant tunnelling results in a highly nonlinear I – V curve and a large bandwidth and these attributes can be employed in a variety of ways to enable applications that include microwave-photonic systems for signal processing [1, 2], secure optical chaos communications [3, 4], and neuromorphic (brain-inspired) photonic applications [5]. Neuromorphic applications and tunnelling devices have a long association that extends back to 1962 and the work of Nagumo with the original Esaki diodes [22] that eventually evolved into the Fitzhugh–Nagumo model of the neuron [23, 24].

The RTD and semiconductor device tunnelling phenomena generally continue to be a rich source of experimental examples of nonlinear dynamics concepts that promise a large range of practical applications.

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Resonant Tunneling Diode Photonics

Devices and applications

Charlie Ironside, Bruno Romeira and José Figueiredo

Chapter 7

Resonant tunnelling diode—optoelectronic oscillator (RTD—OEO)

7.1 Introduction

In the past decade there has been intensive research into novel photonic-based oscillator technologies capable of generating microwave carriers in both electrical and optical domains. Photonic-based oscillators have the potential to replace existing microwave resonator-based oscillator technologies since these cannot match all advantages of photonic systems. In a practical oscillator for photonic applications, it is possible to synchronise or control the oscillator using both electrical and optical signals. There are a wide-range of methods of generating a microwave signal in the optical domain. In figure 7.1, a few examples are given of the most studied, based on either direct or external modulation of a semiconductor laser diode. The direct modulation example, figure 7.1(a), corresponds to one of the lowest cost solutions. The external modulation, figure 7.1(b), uses a laser, an external electro-optic modulator and a stable microwave signal from a local oscillator. This method provides higher performance with a trade-off between cost and energy consumption.

The generated microwave carriers using either direct or external modulation of a semiconductor laser can be substantially improved for higher performance using either injection locking or delayed feedback techniques, as the methods include optical injection [1], optical feedback [2, 3], and optoelectronic feedback [4, 5]. In the case of the optical injection scheme, figure 7.1(c), an external coherent light source in a master–slave configuration is used to provide bandwidth enhancement, chirp and noise reduction for optical communication applications [6]. In the case of optical feedback (not shown), an external mirror is used to optically re-inject a fraction of the light emitted by the laser back to its active region. The round-trip time of light propagation in the external cavity introduces an optical delay in the system. In the case of optoelectronic feedback, a fast responding photodiode is used to detect the light output followed by current re-injection to the RF input of the laser. This

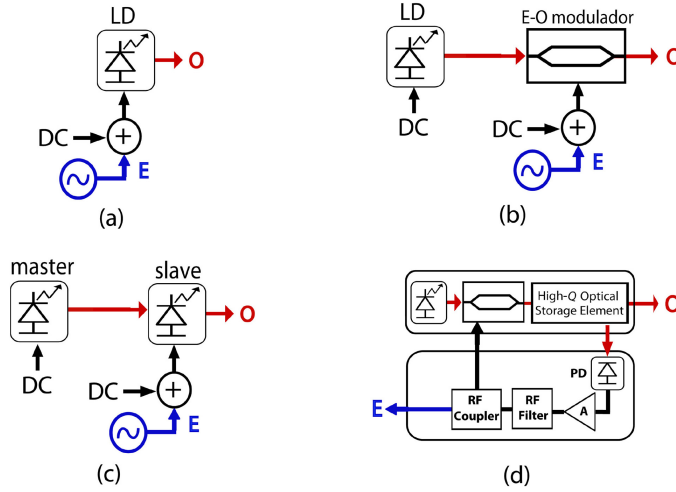


Figure 7.1. Methods of generating microwave electrical (E) signals in the optical (O) domain. (a) Direct modulation. (b) External modulation. (c) Optical injection locking of a directly modulated laser. (d) The optoelectronic oscillator (PD: photo-detector; A: amplifier).

generates a modulated current signal that is an image of the laser output which is added, after an optoelectronic delay time, to the laser injection current.

A unique example of the implementation of the optoelectronic feedback method is the optoelectronic oscillator, figure 7.1(d). In this case, the laser diode is used only as a continuous wave light source, whereas external electronic and/or electro-optic components (e.g. electro-optic modulators, RF oscillators, etc) are used as the nonlinear elements. During the past two decades, several examples of time-delayed OEO topologies combining electronic and photonic components have been intensively reported showing the most promising results for the generation of stable and high-purity RF signals and these will be the focus of this chapter.

In this chapter, after a general introduction of OEO systems, a description of the experimental arrangement of an optoelectronic oscillator fibre-based architecture using self-synchronised resonant tunnelling diode photonic integrated devices is presented, followed by the key results recently achieved operating this architecture in either single or dual loop fibre configurations. The experimental results are compared with numerical simulations of a time-delayed feedback Liénard oscillator model. Lastly, examples of applications using time-delayed feedback RTD—OEOs are given, including clock generation and optical data buffering. This chapter is adapted from chapter 5 of [7]; it has been extensively revised and updated.

7.2 Optoelectronic oscillator (OEO)

An OEO is a feedback oscillator in a hybrid electro-optical configuration which converts optical energy from a laser to microwave signals. In place of a resonant circuit, the OEO uses a long and low-loss fibre-optic delay line as the energy storage element. In 1996, at the Jet Propulsion Laboratory, the first highly stable OEO

system was demonstrated (for a perspective review, see [8]). It consists of an optical fibre delay line several kilometres long, a wide-band electro-optical modulator, light from a semiconductor laser coherent light beam, a photodiode, a narrow microwave RF filter, and microwave and/or optical amplifiers. Its basic architecture is shown in figure 7.1(d) and it has been one of most studied configurations. The conditions for self-sustained oscillations include the coherent addition of partial waves each way around the loop (the loop phase shift must be an integer multiple of 2π), and a loop gain exceeding losses for the circulating waves in the loop. These form the Barkhausen criteria [9] to obtain stable oscillations in the OEO. Because of the extremely low losses and large energy storage associated with the fibre, a high loaded resonator quality factor can be obtained which results in oscillations with extremely low-phase noise. Therefore, OEOs have interesting applications in optical communication links and in precise test and measurement equipment.

Several alternative OEO topologies to the one presented in figure 7.1(d) showing stable and low-phase noise RF carriers have been reported in the last few years using either direct or external modulation of laser diodes [10], or heterojunction bipolar transistors [11], to name a few. However, in order to obtain RF signals in both electrical and optical domains they require several additional electronic and optoelectronic components such as phase-locked local oscillators, amplifiers, photo-detectors, light emitters and/or amplifiers or modulators. Furthermore, techniques including bidirectional injection or self-injection locking may be required to achieve the very low-phase noise levels needed by some applications. Therefore, these OEO alternatives are rather complex and costly for applications such as radio-over-fibre networks where compact solutions for the transmission of RF signals are needed [12, 13].

Although the choice of the laser source can have an effect on the performance of the OEO (specifically, with respect to the impact of laser intensity noise on the output phase noise), most conventional schemes require the use of an external modulator (normally Mach–Zehnder) to introduce a microwave sub-carrier onto the optical signal. This places a practical limit on the frequency of oscillation, based on the bandwidth of the electro-optic modulator. Importantly, to obtain a highly stable and single oscillation signal it is necessary to include a filter in the feedback loop to remove the harmonic signals generated by the nonlinear response of the electro-optic modulator.

To address the limitations imposed by OEOs, namely mode selection filtering, various approaches have been proposed to select a single operational mode from those available. The most promising implementations use photonic generators via optical frequency division [14], figure 7.2(a), and laser self-injection locking using optical feedback from high- Q whispering gallery mode (WGM) resonators [15], figure 7.2(b). These methods provide ultra-stable electrical signals comparable to those produced by the best microwave oscillators and clearly have the advantages of using optical methods for the generation of high-frequency and low-phase noise microwave carriers. Figure 7.3 summarises the single-sideband (SSB) phase noise for several leading microwave generation technologies in the 10 GHz range.

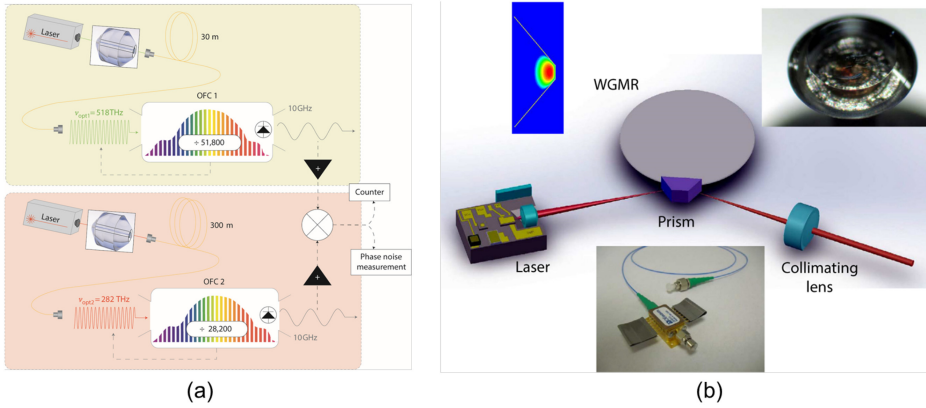


Figure 7.2. (a) Schematic of the experimental set-up used for generation and characterisation of the 10 GHz low-noise microwaves via optical frequency division¹⁹. (b) Ultralow noise miniature external cavity semiconductor laser²⁰. Light from the pump laser semiconductor DFB laser enters the whispering gallery mode resonator (WGMR) through the prism. Part of light is reflected back to the laser because of Rayleigh scattering in the resonator. The light exiting the prism is collimated and used for processing. Insets show power distribution in the resonator mode (top-left), a picture of the actual resonator (top-right) as well as a picture of the packaged laser (bottom).

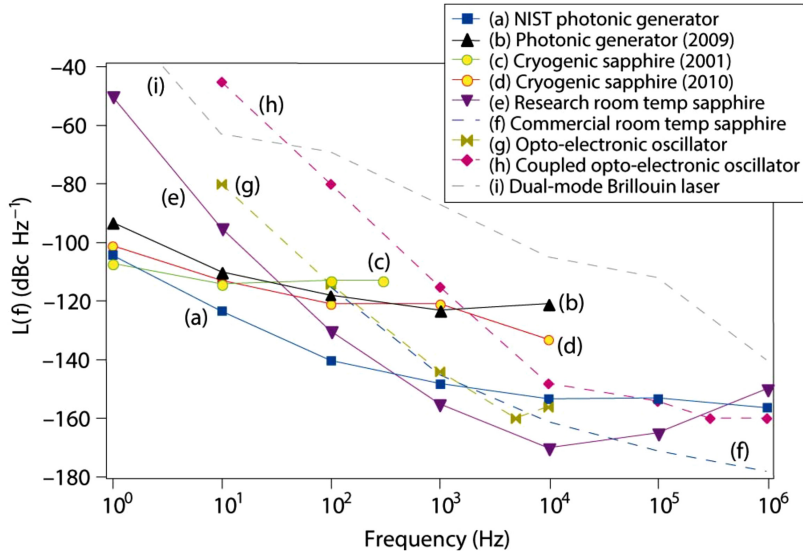


Figure 7.3. Single-sideband phase noise for several leading microwave generation technologies in the 10 GHz range¹⁹. (a) Result of the NIST photonic generator (b) Previous Er:fibre and Ti:sapphire optical frequency divider results. (c),(d) Cryogenic sapphire oscillators. (e) Research room-temperature sapphire oscillator (f). Commercial room-temperature sapphire oscillator. (g) and (h) Long fibre (g) and coupled (h) optoelectronic oscillators. (i) Upper limit of dual-mode Brillouin laser.

Although remarkable results have been achieved, most of the aforementioned methods are complex, costly, and require a large number of electronic and optical high-bandwidth amplification components which consume large amounts of energy for amplification purposes and are also major noise contributor elements. Thus, a great deal of work on OEOs has been developed to provide high-performance and high-power microwave and millimetre-wavelength oscillators that not only are compact but also have power consumption many orders of magnitude lower than existing devices.

In what follows, our recent work on an optoelectronic oscillator fibre-based architecture using self-synchronised resonant tunnelling diode photonic integrated devices is discussed, offering a new and alternative solution for compact photonic integrated circuit-based OEOs.

7.3 Photonic integrated RTD—OEO

RTD-based oscillators have typically large noise levels, with typical single-sideband noise power densities larger than -100 dBc Hz^{-1} at 100 kHz offset from the carrier frequency. This compares with the typical noise levels of other analogous electronic free-running oscillators presented recently in [11]. In order to obtain further improvements in the phase noise, external injection locking techniques are usually employed, as discussed in chapter 6 of this book. The RTD—PD based OEO topology discussed here is a robust method of providing stability of the free-running oscillations and further reduce fluctuations and noise without employing bulky external reference sources.

In this section, we discuss the feasibility of using RTD—PD oscillators for low powered photonic integrated OEO systems that considerably simplifies the typical reported OEO configurations reviewed in the previous section and shown in figure 7.4(a). Figure 7.4(b) shows the architecture of a photonic integrated RTD—OEO. It consists of an RTD—PD oscillator operating at multi-gigahertz frequencies. The generated microwave signals directly modulate a laser diode. High-purity microwave signals are enabled by a low-loss high- Q cavity provided by the optical fibre delay line implemented in a feedback loop configuration. The RTD—OEO does

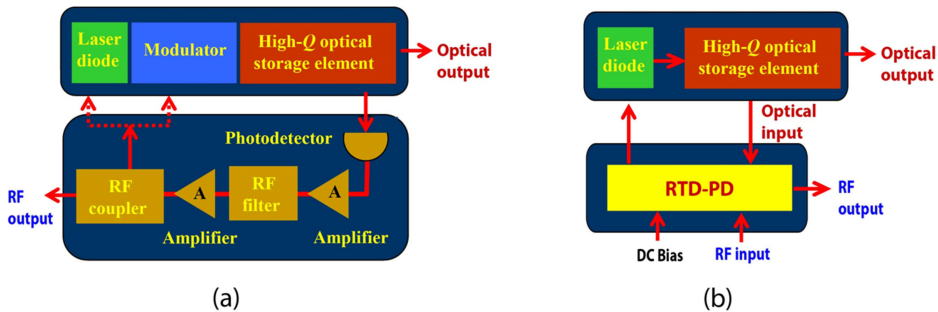


Figure 7.4. (a) First reported OEO architecture of an optoelectronic oscillator. (b) A self-synchronised OEO based on an RTD—PD driving a laser diode on time-delayed feedback implemented using a low-loss high- Q cavity.

not require low-noise and stable external RF amplifiers nor microwave filters usually required to compensate E/O and O/E conversion losses and frequency mode selection, respectively. By eliminating the RF and EDFA amplifiers from the OEO, the main elements contributing to the noise of an OEO are eliminated. Moreover, the RTD—OEO has a low power requirement compared to previous OEOs and no further high-speed, high-cost, and bulky components such as photo-detectors or modulators are needed. In what follows, two main architectures are presented (i) the single-optical fibre loop, and (ii) the dual-optical fibre loop configuration. The stability and phase noise characteristics as a function of the optical power and the length of fibre delay line are discussed.

7.3.1 Single-optical fibre loop

The single-optical fibre loop RTD—OEO is realised in a closed loop configuration, figure 7.5(a), consisting of a delay line implemented using a single-mode optical fibre roll, a DBQW—RTD monolithic waveguide photo-detector oscillator, and a commercial laser diode source. With this simple feedback route configuration, we obtained an RTD—OEO having both electrical and optical input and output ports, with the RTD—OEO frequency and phase noise fluctuations substantially reduced after passing the high- Q optical delay line. The utilisation of fibre-rolls with different lengths provide an easy way of varying the optical feedback route of the RTD—OEO, but other more compact delay lines could be considered, including whispering gallery mode resonators.

In the RTD—OEO analysed here, both the RTD—PD and the LD are hybrid integrated on a printed circuit board transmission line layout, as described in chapters 5 and 6. The RTD—PD devices consist of a chip die with a size of around $0.5 \times 0.4 \text{ mm}^2$ (see chapter 4 for details), The typical optical responsivity of the detector at around $1.55 \text{ }\mu\text{m}$ was $\sim 0.28 \text{ A W}^{-1}$. These values were measured when the RTD—PD was DC biased around 2.4 V , i.e., when operating close to the valley region, figure 7.5(b). Electro-optical conversion of the microwave signals is achieved by direct modulation of a laser diode operating around $1.55 \text{ }\mu\text{m}$ and with an

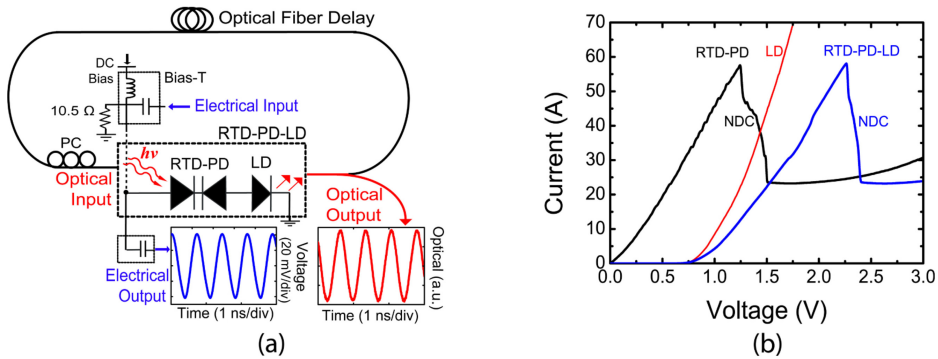


Figure 7.5. (a) Schematic of the self-synchronised RTD—OEO set-up. (b) I - V characteristics of the RTD—PD—LD, and of the RTD—PD and LD devices showing the negative differential conductance (NDC) region.

efficiency of $\sim 0.25 \text{ W A}^{-1}$. As shown in figure 7.5(a), this arrangement provides a simple configuration without requiring external signal amplification. However, because the RTD—PD devices were not optimised for this particular experiment, in some experiments, namely in the dual loop fibre experiments presented in the next section, EDFA amplifiers were used to compensate the losses of coupling the light in more than one optical fibre.

The RTD—OEO operates as a self-synchronised optoelectronic voltage controlled oscillator, that is, the free-running frequency of the RTD—OEO oscillator is controlled by adjusting the DC bias voltage. When biased in the NDC region, free-running electrical current oscillations build up, directly modulating the LD. The LD optical modulated output is then coupled to a lensed single-mode fibre, sent through an optical fibre delay line, and finally re-injected into the RTD—PD waveguide optical input using an identical single-mode lensed fibre. The optical re-injection mechanism enables self-injection locking of the RTD free-running oscillations. The operation frequency of the RTD—OEO is determined by the RTD—PD parallel capacitance and the series equivalent inductance of the gold wires used to connect the RTD—PD to the LD with typical lengths of $\sim 4 \text{ mm}$, enabling us to estimate an equivalent inductance of 3.6 nH . For the photonic integrated circuit discussed here, a frequency operation of up to 1.4 GHz could be achieved. Using an optimised circuit configuration for layouts that significantly reduce the series inductance, the oscillator could operate $>10 \text{ GHz}$. Since the laser used had a cut-off 3 dB frequency of 11 GHz , the laser is the device that limits the maximum frequency bandwidth that can be achieved with this OEO configuration. In order to achieve frequencies well above the maximum bandwidth available in laser components ($>30 \text{ GHz}$), one can replace the laser by an RTD—EAM device, discussed in chapter 3, or by other external modulator devices.

In the configuration shown in figure 7.5(a), without optical re-injection, the RTD—OEO oscillates with a free-running frequency tuneable from 1.05 to 1.41 GHz , depending on the DC bias voltage. When biased close to the valley region, $\sim 2.4 \text{ V}$, the frequency of the electrical and optical modulated free-running oscillations was around 1.4052 GHz (the oscillations in time domain are shown in the inset of figure 7.5(a)) with an estimated power consumption of 0.55 W . The power consumption can also be further reduced below 100 mW by inserting a capacitor in series with the shunt resistor shown in figure 7.5(a). Figure 7.6(a) presents the free-running oscillation spectrum trace, showing the typical broad spectrum caused by the noise-induced fluctuations of the fundamental frequency oscillation. The measured electrical power is determined by the RTD circuit, the I — V characteristic and the electrical losses of the measurement set-up.

The RTD—OEO performance was investigated using fibre delay lines with lengths, L_f , between 0.4 and 4.4 km . In figures 7.6(b) and (c), we show the spectra of the self-synchronised free-running oscillations when two fibre roll lengths of 0.8 km and 1.2 km , respectively, are used. The average in-fibre optical power was $P \sim 5 \text{ dBm}$. Also presented as an inset at the top of each figure is the spectrum surveillance showing the plot density spectrum. The results show that the self-injection

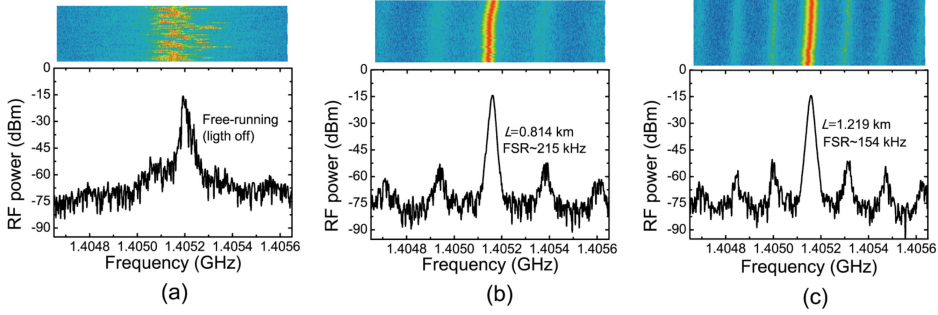


Figure 7.6. RF power spectrum traces of the RTD—OEO output. (a) Free-running. Self-synchronisation at $P \sim 5$ dBm using fibre loops of (b) $L_f = 0.814$ km, and (c) $L_f = 1.219$ km. The frequency span and resolution bandwidth settings for all figures were 1 MHz and 10 kHz, respectively. The top part of each figure shows a spectrogram that represents the evolution of spectral density recorded over 30 s (time in the vertical axis).

enhances the signal quality and improves the frequency stability with a substantial reduction of the phase noise.

The side modes presented in figures 7.6(b) and (c) are separated from the centre frequency by about 215 kHz and 154 kHz, respectively, with a single-mode suppression ratio (SMSR) of 38 dB and 36 dB, respectively. These correspond to the mode spacing of the free spectral range (FSR) of the RTD—OEO and are related to the time delay τ_d , which in this case depends on the electrical time delay τ_e , due to the electrical components, and the optical time delay τ_{op} , due to the length of the optical fibre. In the experiments discussed here the optical time delay τ_{op} is much larger than the electrical time delay τ_e . Therefore, the FSR of the RTD—OEO can be calculated as:

$$\text{FSR} = \frac{1}{\tau_e + \tau_{op}} \cong \frac{1}{\tau_{op}} \cong \frac{c}{n_F L_f} \quad (7.1)$$

where $\tau_{op} = n_F L_f / c$, with n_F being the optical fibre effective refractive index, and c the velocity of light, that is, the RTD—OEO mode spacing is mainly determined by the fibre length. From equation (7.1) the following FSR values are estimated: $L_f = 0.814$ km, FSR = 251 kHz; and $L_f = 1.219$ km, FSR = 167.6 kHz, which compare with the experimental results shown in figures 7.6(b) and (c), respectively.

In order to verify the phase noise reduction performance, the respective SSB phase noises at the output signals were measured and are shown in figure 7.7. For an in-fibre optical power of $P \sim 9$ dBm and optical fibre lengths of 0.409 km, 0.814 km, and 1.219 km, the corresponding self-injection locked phase noise values are -89 dBc Hz^{-1} , -96.7 dBc Hz^{-1} , and -101.8 dBc Hz^{-1} , respectively, figure 7.7(a). A phase noise reduction more than 40 dB at 10 kHz offset is achieved when compared to the free-running oscillations without light re-injection. In the self-synchronised mode the SSB phase noise at 10 kHz offset from the centre frequency is substantially reduced with the increase of the in-fibre optical injection power and also the fibre delay, as summarized in figure 7.7(b).

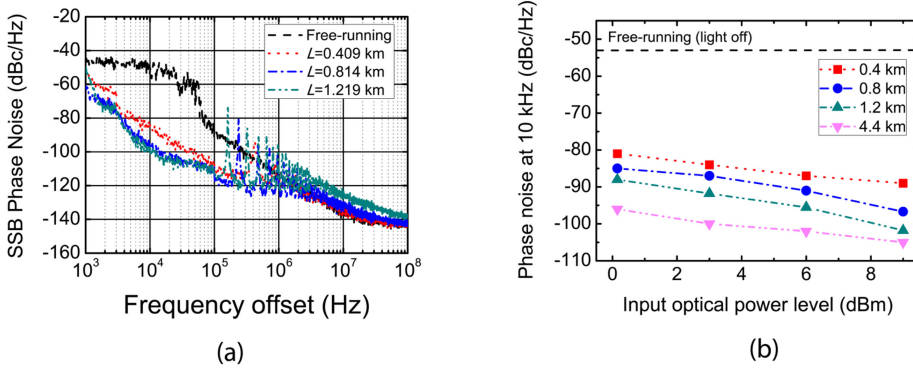


Figure 7.7. (a) Measured SSB phase noises as a function of fiber length at $P \sim 9$ dBm. Also shown in black is the phase noise trace for the free running oscillation. (b) RTD—OEO SSB phase noise measurements at 10 kHz offset from the centre frequency as a function of in-fibre optical power.

In figure 7.7(b) are also presented the results of SSB phase noise for a very long delay line, i.e., using a long fibre roll with $L_f = 4.4$ km. As compared to the free-running oscillations without light re-injection, a phase noise reduction of up to 50 dB is achieved at frequencies offsets below the side modes. Phase noise values below -100 dBc Hz $^{-1}$ at 10 kHz offset are achieved at power levels close to 3 dBm, which corresponds to ~ 17 dB improvement when compared with $L_f = 0.4$ km. In this RTD—OEO, although as the re-injected power level increases the optical losses are lower, the number of side modes are also strongly enhanced. As discussed next, in order to maintain the low-phase noise levels and frequency stability due to the very long delay line, and at the same time suppress side mode signals that deteriorate the performance of the RTD—OEO, the introduction of a multi-loop fibre configuration provides an excellent method for mode selection.

7.3.2 Dual-optical fibre loop

As verified in the previous subsection, a disadvantage of using a single-fibre loop is the appearance of side mode oscillations. These modes—highly undesirable for certain applications—are caused by the propagation of waves multiple times around the OEO loop. There are two effective solutions for the removal, or suppression, of side modes. One scheme involves using multiple loops of fibre, which essentially functions as a narrow band filter. Another technique is to use a high- Q optical cavity to filter out the unwanted modes.

Yao and Maleki demonstrated a dual loop OEO configuration where the optical power was split by an optical decoupler with both portions of the optical signal being subsequently detected by two photo-detectors [16]. The two converted RF signals were coupled by an RF coupler and then fed to the modulator after amplification. In each loop, the process was equivalent to a single-loop OEO. In this configuration, the method uses the natural structure of the OEO cavity to obtain filtering through an additional fibre loop or loops having shorter lengths, in parallel with the longer fibre loop. Under the cooperation of the two loops, the side modes

are efficiently suppressed due to the gain competition between the two sets of modes. In this case, there is no random interference and beat noise because the RF signals have the same oscillation frequency and are coupled in the electrical domain. Although this design gave the expected result, two high-speed photo-detectors were needed.

Here, substantial improvements in the reduction of the level of the side modes using a carrier suppression scheme in a dual loop RTD—OEO configuration is discussed. In this configuration, a portion of the laser optical output is split into two parts each propagating through fibres having different lengths and then combined in a single fibre to be coupled to the single RTD—PD. That is, the RF signals are coupled in the optical domain without additional active electrical devices. Due to the interference between the two combined optical signals, the dual loop RTD—OEO enables a strong mode selectivity with suppression of the side modes. The RTD—OEO circuit configuration and components used in the dual loop experiment were similar to the one presented for the single-fibre experiment. Depending on the DC bias, the circuit under analysis here generated oscillations with natural frequencies ranging from 0.944 GHz to 1.129 GHz. In this experiment, the RTD—OEO was the DC biased at 2.33 V (close to the valley), showing oscillations around 1.121 GHz. The typical optical responsivity of the RTD—PD device was 0.25 A W^{-1} at 1550 nm when biased in the valley region.

The effect of the dual loop RTD—OEO using 4.4 km and 0.4 km fibre loop lengths—see figure 7.8, was studied experimentally measuring the phase noise and level of side modes characteristics. The longer fibre leads to high spectral purity and low-phase noise at low offsets. The short fibre is able to suppress the side modes close to the carrier frequency.

On breaking either path, there is a complete single-loop OEO, which is able to freely run, similar to what was demonstrated in the previous subsection. With the long path broken (4.4 km fibre loop), the mode spacing is 425 kHz and the measured side mode suppression ratio was -53 dBc at $P \sim 6 \text{ dBm}$ —see figure 7.8(a). With the short path broken (0.4 km fibre loop), the mode spacing is 45 kHz and the measured side mode suppression ratio was around -20 dBc —see figure 7.8(b). In the case of

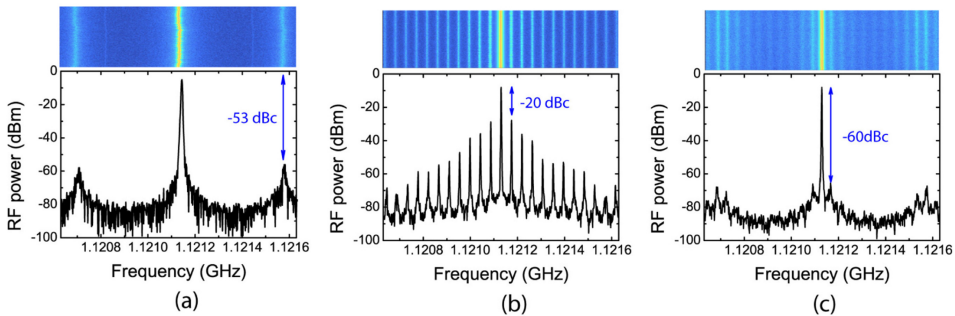


Figure 7.8. (a) RF power spectra output showing the SMSR for (a) 0.4 km single-loop (b) 4.4 km single-loop, and (c) dual loop configuration. The top part of each figure shows a spectrogram that represents the evolution of spectral density recorded over 1 min (time in the vertical axis).

the dual loop configuration, the side mode suppression ratio was improved up to -60 dBc—see figure 7.8(c).

Compared with the single-loop results, the side modes can be effectively suppressed by 20–40 dB, depending on the optical power input launched into the RTD—PD.

The SSB phase noises of the single and dual loop configurations were measured. The phase noise at 10 kHz away from the carrier of the dual loop RTD—OEO was -99.28 dBc Hz $^{-1}$ at 6 dBm in-fibre re-injected optical power, and -102.88 dBc Hz $^{-1}$ at 9 dBm in-fibre re-injected optical power, which compares with the single-loop at 4.4 km, showing -101.67 dBc Hz $^{-1}$ at 10 kHz away from the carrier. Therefore, besides maintaining the quality of the phase noise below -100 dBc Hz $^{-1}$ at low offset, successful suppression of the side modes was achieved in the spectral region above 45 kHz frequency offset of the carrier frequency which confirms the efficient side mode suppression capability of the dual loop configuration. The measurements also showed that additional close-to-carrier noise is introduced by the dual loop configuration resulting in slightly larger noise levels at lower carrier offsets when compared with the single 4.4 km long path.

As a final note, the dual loop OEO implementation discussed here uses a single RTD—PD only, which considerably reduces the number of components usually required to form a dual loop OEO configuration. However, it may suffer from problems related to coupling in the optical domain, namely interference and beating. This occurs mainly because the RF signals are coupled in the optical domain instead of other configurations where the coupling is achieved electrically. In order to avoid these two effects, and retain the function of the dual loop, an effective and simple way is to utilise, for example, a polarisation beam splitter and combiner. The fluctuation of polarisation and loop length are also additional difficulties in the implementation of the RTD—OEO dual loop scheme. If the polarisation has large deviations, the optimised power distribution among the fibres is destroyed, and the side mode suppression ratio is degraded. To eliminate this effect, polarisation-maintaining fibres can be used in the system. In the case of the loop length fluctuations, which are usually caused by environmental changes, these effects are difficult to overcome unless temperature control or feedback control is employed.

In summary, the results of both single and dual loop schemes demonstrate the self-synchronisation substantially improves the phase quality of free-running output signals with a compromise between the length of the time-delayed feedback loop configuration and the phase noise suppression. Although in terms of phase noise characteristics the RTD—OEO is still below commercially available voltage-controlled modules, such as coaxial resonator oscillators frequencies that typically show phase noise values at 1 GHz of around -120 dBc Hz $^{-1}$ at 10 kHz offset, it is expected that further techniques including temperature and vibration control can have a great impact on the RTD—OEO performance. Noteworthy, the results presented here were strongly limited by the responsivity of non-optimised RTD—PDs which can be further be improved by using anti-reflection coating facets and maximising waveguide external quantum efficiency. Lastly, since there is still considerable scope for optimisation in many aspects of the OEO configuration,

namely monolithic integration of the RTD—PD with LD with an on-chip high- Q feedback route, the OEO topologies using self-injection locking of RTD oscillators can be potential alternatives for novel low-cost and on-chip OEOs with a small footprint and with energy consumptions substantially lower than existing devices.

7.3.3 Time-delayed feedback Liénard oscillator model

In chapters 5 and 6 where the theory of RTD and LD dynamical systems was discussed, we have considered systems that can be analysed by a set of first-order ordinary differential equations. The RTD—OEO system belongs to a different class of systems, delay systems. Time-delayed feedback systems, where the dynamic characteristics of a system at a given instant t depend on the state of the system at an anterior time $t - \tau_d$, where τ_d is the time delay, are common in scientific fields such as physiology, biology, epidemiology, and control systems [17]. The effect of the delayed feedback is to re-inject into the system one or more state variables (e.g. current or intensity), with a delay. Depending on the time delay and feedback strength, a delayed feedback loop can be designed to (i) achieve highly stable periodic oscillations, or (ii) exceedingly complex dynamical phenomena and aperiodic oscillations, namely chaos [18, 19]. Notably, the dynamical characteristics of such systems can be easily synchronised and controlled via external electro-optical perturbations [20, 21]. The time-delayed feedback RTD—OEO system is modelled be a system of first-order delay-differential equations:

$$\frac{dx(t)}{dt} = F(x(t), x(t + \tau_d); p) \quad (7.2)$$

where τ_d is the delay of the system and F is the vector field. The derivative of $x(t)$ at time t depends on the value of $x(t)$ at time t and on the value that $x(t)$ took in the past at time $t + \tau_d$. This means that it is necessary to know the vectorial function on the time interval of length τ_d to determine the state of the system and its evolution, because the phase space is an infinite-dimensional functional space. The set of parameters $p = (p_1; p_2; \dots)$ describe the features of a particular system modelled by the vector field. When the vector field depends explicitly on time, the system is said to be non-autonomous.

In this section, we discuss a numerical model of the single-loop delayed feedback RTD—OEO system using DDEs [22]. The model includes a time-delayed feedback Liénard oscillator that models the RTD electrical dynamics and the feedback loop of the OEO, laser rate equations describing the LD dynamics, and white Gaussian input noise to describe the dominant OEO noise sources, the thermal and shot noise.

The dynamics of the RTD—OEO is analysed considering the lumped electrical circuit of figure 7.9. This circuit is equivalent to a Liénard oscillator [23, 24], that includes photocurrent (I_{ph}) and current noise (I_n) terms. The RTD—PD is represented by its intrinsic capacitance in parallel with a voltage dependent current source $F(V)$, and its photo-detection characteristics are modelled as an optical power dependent current source, I_{ph} [7] (see chapter 4 for more details on the photo-detector).

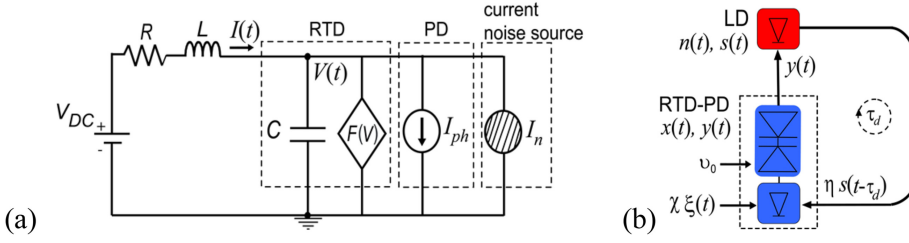


Figure 7.9. (a) Equivalent electrical circuit schematic of the resonant tunnelling diode photo-detector oscillator with current noise injection. The resistor R and inductor L account to the circuit series and parasitic resistance and inductance of RTD—PD—LD circuit, respectively. (b) Block diagram of the mathematical representation of the delayed feedback Liénard oscillator laser diode dimensionless system.

Using the Kirchhoff's rules, the same procedure discussed in chapter 6 was applied to obtain the two dimensionless first-order differential equations that describe the electrical circuit of figure 7.9(a). Then, the electrical system was coupled to the laser diode rate equations that model the laser system. As previously, time is normalised to the characteristic frequency of oscillation of the solitary RTD—PD—LD, $\omega_0 = (\sqrt{LC})^{-1}$, hence $t = (\omega_0)^{-1}\tau'$. Finally, redefining τ' as t and introducing into the system the optical delayed feedback $s(t - \tau'_d)$, η is the feedback strength and τ'_d is the time delay with respect to the dimensionless time t , the Liénard oscillator-laser diode dynamical system is transformed into the following dimensionless coupled DDEs:

$$\frac{dx(t)}{dt} = \frac{1}{\mu} [y(t) - f(x) - \chi \xi(t) - \eta s(t - \tau'_d)] \quad (7.3)$$

$$\frac{dy(t)}{dt} = \mu [\nu_0 + \gamma y(t) - x(t)] \quad (7.4)$$

$$\frac{dn(t)}{dt} = \frac{1}{\tau'_n} \left[\frac{i_m(t)}{i_{th}} - n(t) - \frac{n(t) - \delta}{1 - \delta} \{1 - \epsilon s(t)\} s(t) \right] \quad (7.5)$$

$$\frac{ds(t)}{dt} = \frac{1}{\tau'_p} \left[\frac{n(t) - \delta}{1 - \delta} \{1 - \epsilon s(t)\} s(t) - s(t) + \beta n(t) \right] \quad (7.6)$$

The above equations describe the optical feedback control of the RTD—PD—LD through the variable $s(t - \tau'_d)$. The feedback strength η parameter depends on the fraction of the laser optical output power introduced by the delayed feedback loop and on the RTD—PD detection characteristics. The function $f(x)$ comes from the normalisation of $F(V)$, and $\mu = V_0/I_0\sqrt{C/L}$ is dimensionless. Equations (7.5) and (7.6) are the dimensionless rate equations describing LD normalised photon $s(t)$ and injected carrier $n(t)$ densities, respectively, where $\delta = \frac{N_0}{N_{th}}$ and $\epsilon = \epsilon_n S_0$ are two dimensionless parameters (where N_0 is the carrier density for transparency, N_{th} is the carrier density at threshold, ϵ_n is the value for the nonlinear gain compression factor,

and $S_0 = \Gamma(\tau_p/\tau_n)N_{th}$ where Γ is the optical confinement factor), $i_m(t)$ accounts for the bias current and the current oscillations produced by RTD—PD electrical model, equation 7.5, and i_{th} is the dimensionless laser diode threshold current. The parameters τ_p' and τ_n' come from the time rescaling.

Figure 7.9(b) presents a block diagram of the mathematical representation of the delayed feedback system and respective variables and key parameters.

In real systems, the unavoidable noise sources affect the dynamics by introducing amplitude and period fluctuations even in the most stable periodic signals. In our case, several sources or random processes are at play, like, e.g. thermal and shot noises. We model their overall effect as an effective delta-correlated Gaussian white noise of zero mean input source $\chi\xi(t)$, equation (7.3), where the parameter χ is the dimensionless variance of the distribution and denote the noise strength.

In what follows, we compare the delayed feedback Liénard OEO model in the single-loop configuration with experimental results. When the time delayed feedback is included, important dynamical effects are observed that include close-to-carrier noise reduction and the appearance of side modes due to the delay contribution, similar to the experimental results reported previously in the single-loop configuration.

Figure 7.10 presents the experimental (a) and simulated (b) power spectra around the fundamental free-running oscillation with and without delayed feedback. Figure 7.10(a) (blue trace) presents the experimental results of self-synchronised electrical output using a in-fibre optical re-injected power $P \sim 6$ dBm and 0.4 km fibre length. The results show phase noise and linewidth reductions at offsets below 250 kHz of the carrier frequency, and side modes separated by about 425 kHz with a SMSR of -43 dBc. Note that the linewidth of the self-synchronised output could not be resolved by the 3 kHz resolution limit set by the instrument. Figure 7.10(b) (blue trace) shows the corresponding simulation for a time delay of $\tau_d = 2.35$ μ s, and a feedback strength to noise ratio of $\theta = 2.0$ (corresponding to a feedback strength

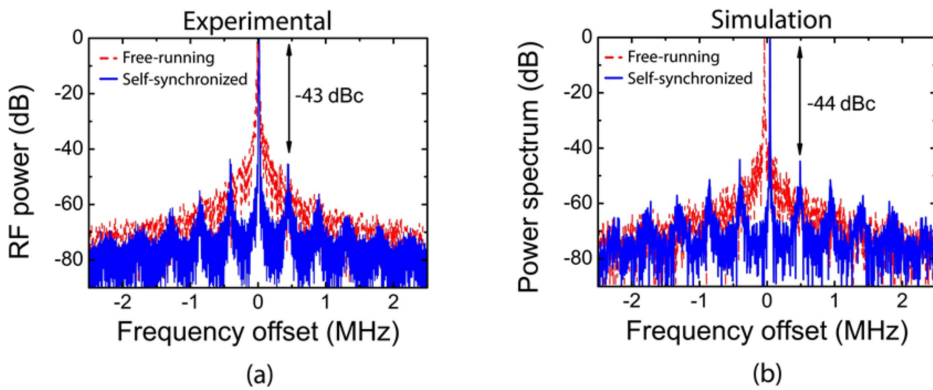


Figure 7.10. (a) Experimental RF power spectra of free-running oscillation (red) and self-synchronised electrical output (blue) and $P \sim 6$ dBm and 0.4 km fibre length. (b) Simulated power spectra of free-running fundamental oscillation and self-synchronised $x(t)$ outputs with time delay of $\tau_d = 2.35$ μ s and feedback strength to noise ratio of $\theta = 2.0$. In both plots the span was 5 MHz and the central frequency was 1.122 07 GHz.

$\eta = 1 \times 10^{-3}$, for an injected optical power of 6 dBm). As observed experimentally, the introduction of a time delayed feedback variable in the Liénard OEO model narrows the linewidth of the fundamental oscillation frequency and generates frequency side peaks due to the time delay, which corresponds to the FSR of the RTD—OEO system. The SMSR of -44 dBc compares with the experimental result shown in figure 7.10(b).

The dynamical properties of the OEO model investigated here follow the dynamics observed experimentally where the characteristics of the OEO can be controlled using optical delay lines. The large number of side-bands spaced by FSR (inversely proportional to the time delay) is an indication that more complex dynamics may occur, which is strongly dependent on the feedback level and the length of the external cavity. In the results presented here, considering the moderate levels of feedback employed, $\eta \ll \Delta I/I_0$, where ΔI is the peak-to-valley current ratio, only stable self-synchronised oscillations were observed. The model can be used either to map the dynamical regimes of RTD—OEOs in order to avoid unstable signals or to analyze the rich dynamical characteristics which can be exploited for innovative applications such as generation of frequency combs [5], or even chaotic transitions [25] controlled by the delayed feedback parameter.

7.4 Regenerative RTD—OEO

Novel methods of buffering of optical signals are highly desirable to avoid congestion of information traffic and to realise efficient optical interconnects. Recently, all-optical buffer memories have been proposed based on slow-optical delay lines or using the Kerr nonlinearity on a standard silica optical fibre providing functionalities such as all-optical storage and reshaping. However, for realistic applications, an optical buffer has to be compact for on-chip integration, which rules out most existing schemes, as they are not easily scalable to a millimetre size footprint and the writing process is often complex and costly. Therefore, configurations that combine the robustness of semiconductor nanoelectronics and optical devices with the wide-bandwidth of photonics elements offer significant advantages because they can provide small size, high-speed, and low-cost alternatives to the all-optical buffers currently proposed.

In this section, we discuss the implementation of regenerative photonic memories using the RTD—OEO system described in previous sections. However, in this case the optical delay line with a time delay inserted off-chip provides a mechanism of re-injection of fired excitable pulses, a type of nonlinear pulse generation analysed in detail in chapter 6. The regenerative memory operates under the physical principle of excitable regeneration [26] in which a generated pulse in the RTD—PD—LD recirculates indefinitely in an optical fibre delayed feedback loop enabling robust regenerative signal buffering with potential to implement memory and logic operations.

This regenerative system follows the strong interest that has been devoted during the past two decades on the effects of communication time delays in biological systems and how they can influence the synchronisation dynamics between distant

coupled neurons. For instance, it was shown that dynamical systems mimicking coupled neurons exhibit stable periodic pulsating regimes, instead of a stable steady-state regime. Following the seminal work of Ikeda, the concept of pattern memorisation in time delayed bistable systems was addressed in optoelectronic systems.

Departing from these initial works, our work involves an excitable element, the RTD—PD—LD resonator, as the nonlinear node (see chapter 6 for more details about RTD-based excitable system). Because of this crucial difference, our approach benefits from the robustness and self-healing properties which are typical of neural signals and do not exist in bistable systems. Specifically, the excitable response of the nonlinear node guarantees a strong and well-defined all-or-nothing pulse response, almost identical in shape and duration to any supra-threshold incoming signal, enabling the reshaping and healing functionality of incoming signals which can be employed in various key communication buffer functions which include writing, storage, reshaping/healing, and XOR operation.

Figure 7.11 presents a typical example of experimental time traces showing the writing and storage of the regenerative memory using electrical injection of a complex pattern of 8-bits (11 011 101). The binary-coded data streams were generated using an arbitrary function generator. For the purposes of demonstration and experimental convenience, the data signals were injected electrically although the signals could be also injected optically, taking advantage of the optical input port of the RTD—PD. By operating the RTD—OEO in an excitable regime (close to the NDC region), we are able to store and regenerate optical bits of information in the fibre, the empty region signalling the ‘0’ bits and the excitable optoelectronic pulses the ‘1’. When these bits are re-injected into the RTD—PD, they trigger the

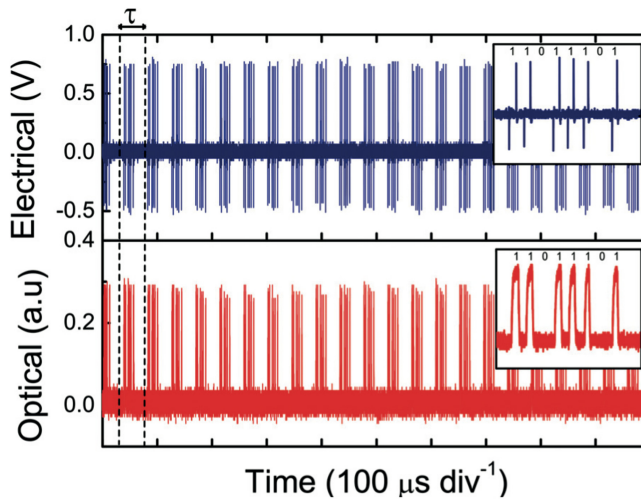


Figure 7.11. Experimental recorded time traces of the regenerative memory showing storage of an 8-bit binary-coded data stream (11 011 101) in a 46 μs cavity round-trip time. The insets show the time traces of a single round-trip.

generation of a new excitable cycle. This regenerative mechanism occurs after each round trip in the fibre which is extremely robust since even if the bit sequence is strongly deteriorated—the all-or-none nonlinear response of the excitable RTD—PD enables the signal regeneration. In the experiments, we used a low-loss optical fibre loop with a $46\ \mu\text{s}$ cavity round-trip time, τ . The time τ was chosen in order that the memory buffer is much larger than the typical excitable lethargic time, T_1 , of the RTD—PD—LD excitable system (in this case, $T_1 \sim 500\ \text{ns}$). The regeneration lasted more than 10^4 round-trips, only limited by the acquisition time of the oscilloscope employed. Real time measurements suggest that storage of a data stream can be achieved for several minutes without using any temperature or vibration controllers of both optical fibre and optoelectronic circuits.

It is noteworthy that the writing process of the regenerative memory can be scalable to multi-gigahertz operation, only limited by the laser diode frequency response. The information can be written with a single addressing pulse, either using electrical or optical incoming data, and can be stored in standard low-loss single-mode silica optical fibres, which is very beneficial for practical applications. In summary, exploiting such a peculiar neuron-inspired excitable electro-optical response using OEO systems is a conceptual breakthrough that has been almost unexplored in the field of optical data communications and could open up new applications, e.g. clock and timing, optical memories and data buffers.

7.5 Conclusions

In conclusion, in this chapter we have discussed an optoelectronic self-synchronised oscillator based on the integration of an RTD—PD for O/E conversion, a laser diode for E/O conversion, and an optical fibre delay line as an optical storage element, described as the RTD—OEO. With this configuration stable free-running self-locked oscillations with a phase noise reduction of more than $>50\ \text{dB}$ at $10\ \text{kHz}$ frequency offset from the carrier frequency were achieved, without requiring RF or optical amplification. When a dual loop topology was introduced, the side modes associated with the delay line could be strongly suppressed. The level of the side modes was suppressed by $40\ \text{dB}$ resulting with overall side mode suppression ratio of $-60\ \text{dBc}$. The RTD—OEO enables the generation of high-purity microwave signals at multi-gigahertz frequencies.

We have also discussed the implementation of a regenerative photonic memory using the RTD—OEO system. The regenerative RTD—OEO memory operates under the physical principle of excitable regeneration in which a generated pulse in the RTD—OEO re-circulates indefinitely in an optical fibre delayed feedback loop enabling robust regenerative signal buffering. Regenerative RTD—OEOs enables the reshaping and healing of incoming electrical or optical signals which can be used to implement key buffer functions required in optical communications such as writing, storage, reshaping/healing, and XOR operation.

Lastly, in this chapter we have introduced a numerical model using delayed differential equations based on a time-delayed feedback Liénard oscillator driven by white Gaussian noise and coupled with laser rate equations. The model predicted the

experimental behaviour of the RTD—OEO as a function of the time delay and the feedback strength. This model, once extended to include all the noise contributions of a typical OEO, such as flicker noise from environmentally driven changes in the fibre length, will enable scientists and engineers to accurately design OEOs based on RTD—PD devices and explore a variety of architectures. Furthermore, the complex phenomena expected in the delayed feedback system can be explored in various applications, not covered in this chapter, namely chaos-controlled methods to stabilise a desired dynamical behaviour [27] and noise-induced stochastic oscillatory behaviour, using the time delayed feedback as a control parameter [28].

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Resonant Tunneling Diode Photonics

Devices and applications

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Chapter 8

Conclusions

This book has presented work on a mesoscopic semiconductor device relying on quantum tunnelling for its operation, the resonant tunnelling diode (RTD); it has concentrated, in particular, on how the double barrier quantum well RTD can be incorporated into photonics with devices such as the RTD—EAM (chapter 3), the RTD—LD (chapter 5) and the RTD—PD (chapter 4). Because the double barrier quantum well RTD consists of just a quantum well sandwiched between two thin barriers it is easily integrated into existing semiconductor optoelectronic components. Further, these RTD enhanced components can be incorporated into systems such as the optoelectronic oscillator (OEO), chapter 7, and neuromorphic computing systems, chapter 6.

Quantum tunnelling is entirely a quantum effect with no classical analogue that can be realised using semiconductor heterostructures and operates at room temperature. The nonlinear current–voltage characteristics associated with quantum tunnelling can be configured to provide negative differential conductance (NDC), and thereby, large electrical gain in high bandwidth (up to ~2 THz) devices. Further, compared to the avalanche photodetector (APD), this gain is relatively low noise. So far, the bandwidth of the RTD gain has been limited not by the tunnelling process but by the components that make up the circuit such as the electrodes for providing the DC bias.

The RTD—EAM, RTD—PD and RTD—LD, all have built-in high bandwidth and low noise gain. A feature that can be utilised to improve existing systems such as optical communications but can also be employed in new system concepts such as the RTD—PD optically controlled 79 GHz RF oscillator for optical to wireless data conversion.

In addition, the high bandwidth with gain means that it is very easy to make the devices oscillate. The oscillation frequencies are determined by the external components and range from a few MHz to up nearly 2 THz. This unique feature

of RTD photonic devices has been exploited in the OEO and for neuromorphic processing and memory.

A brief survey of the history of tunnelling devices reveals that it was realised very early on, and because of the highly nonlinear current–voltage characteristic, they provide an excellent system for the exploration of nonlinear dynamics and the implementation of neuron like behaviour as in the FitzHugh–Nagumo model [1] of nerve operation.

The trend in semiconductor electronics is for ever smaller device features—in 2014 CMOS reached the 14 nm node [2, 3] and as of 2019 it is approaching the 7 nm node. At these scales, the basic physics of mesoscopic semiconductor devices—as outlined in chapter 1—means that it is very likely that an understanding of quantum tunnelling will play an ever more important role as the simple ballistic transport models become more and more inadequate. The width of the barriers in RTD devices is around 2 nm, so RTDs reached the 2 nm node in the 1990s and a great deal of expertise has been accumulated in how to utilise tunnelling effects to improve and extend photonic devices. So, the future of RTD based devices is well aligned with the dominant trend in semiconductor electronics and optoelectronics.

In this book we have covered RTD photonic devices and their application in a variety of systems but more generally RTD photonics has been a major influence in the efforts to use quantum tunnelling in semiconductor devices to improve and extend their performance.

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